IN THE UNITED STATES DISTRICT COURT
FOR THE EASTERN DISTRICT OF TEXAS
SHERMAN DIVISION

MICROTUNE (Texas), L.P.,

Plaintiff,

v.

BROADCOM CORPORATION,

Defendant.

CIVIL ACTION NO. 4:01CV23

DEMAND FOR JURY TRIAL

MICROTUNE (TEXAS), L.P.'S MARKMAN BRIEF

76

I.

INTRODUCTION

Microtune (Texas), L.P. ("Microtune"), located in Plano, Texas, owns United States Patent No. 5,737,035, entitled Highly Integrated Television Tuner On A Single Microcircuit (the "'035 Patent"). The inventions disclosed in the '035 Patent solve several problems with prior art television tuners, including cost, size, and the need for manual adjustment at time of manufacturing. The '035 Patent allows television tuners to be integrated into small electronic devices and within computers, which previously was not possible.

This brief discusses (1) the law regarding claim construction (Section II); (2) relevant technical background to the invention, using graphics to aid the discussion (Section III); and (3) the definitions of claim terms (Section IV). The '035 Patent is attached as Exhibit A. A chart cross-referencing the proposed definitions to the actual claim language is attached as Exhibit B. A list of the claim term definitions agreed to by both parties is attached as Exhibit C. The declaration of Bernard J. Lechner accompanies this document ("Lechner Decl."). Bernard Lechner is an expert electrical engineer in the television field, and his declaration provides the evidentiary support for the technical background section. Microtune relies, however, on intrinsic evidence for claim interpretation.[1]

II.

CLAIM INTERPRETATION

Claim interpretation is a pure question of law. *Cybor Corp. v. FAS Technologies, Inc.*, 138 F.3d 1448, 1456 (Fed. Cir. 1998); *Markman v. Westview Instruments, Inc.*, 52 F.3d 967, 979 (Fed. Cir. 1995) (en banc), *aff'd*, 517 U.S. 370 (1996). In construing claims, the Court should first look to intrinsic evidence, consisting of the claims themselves, the specification and the

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[1] Microtune has asserted claims 83-96, claims 102-115, and claims 120-128 of the '035 patent. Microtune contemplates that additional claims could be added in good faith as discovery proceeds, but this brief covers Microtune's proposed construction of claim terms found in the asserted claims.

prosecution history. *Vitronics Corp. v. Conceptronic, Inc.*, 90 F.3d 1576, 1582 (Fed. Cir. 1996). "Such intrinsic evidence is the most significant source of the legally operative meaning of disputed claim language." *Id.*

Although intrinsic evidence is the most significant source of the legally-operative meaning of disputed claim language, all intrinsic evidence is not equal. *See Vitronics*, 90 F.3d at 1582. "The first requirement in claim interpretation is to examine the claim language." *SmithKline Diagnostics, Inc. v. Helena Laboratories Corp.*, 859 F.2d 878, 882 (Fed. Cir. 1988). The words in the claim are generally given their "ordinary and customary meaning." *Vitronics*, 90 F.3d at 1582. In fact, there is a "heavy presumption in favor of the ordinary meaning of claim language." *Johnson Worldwide Assocs., Inc. v. Zebco Corp.*, 175 F.3d 985, 989 (Fed. Cir. 1999). "Claim terms cannot be narrowed by reference to the written description or prosecution history unless the language of the claims invites reference to those sources." *Id.* at 989-990. Such invitation occurs when the patentee has set forth an explicit definition for a claim term, or when there is no means by which the scope of the claim can be ascertained from the language used. *Id.; see also York Prods., Inc. v. Central Tractor Farm & Family Ctr.*, 99 F.3d 1568, 1572, 40 U.S.P.Q.2d 1619, 1622 (Fed. Cir. 1996) ("Without an express intent to impart a novel meaning to claim terms, an inventor's claim terms take on their ordinary meaning.").

This is not to say that claim terms are to be construed in a vacuum. A word in a patent claim must be interpreted as one of ordinary skill in the art would interpret it unless it is apparent from the patent and prosecution history the inventor used the term with a different meaning. Usually, the specification is the "single best guide" to the meaning of a disputed term and may serve as a dictionary, defining the particular meaning of terms used in the claims. *Vitronics*, 90 F.3d at 1582; *Markman,* 52 F.3d at 979.

> [I]t is always necessary to review the specification to determine whether the inventor has used any terms in a manner inconsistent with their ordinary meaning. The specification acts as a dictionary when it expressly defines terms used in the claims or when it defines terms by implication . . . . Thus, the specification is always highly relevant to the claim construction analysis. Usually, it is dispositive; it is the single best guide to the meaning of a disputed term.

Vitronics, 90 F.3d at 1582.

If, after reviewing all available intrinsic evidence, some genuine ambiguity still exists in the claims, the Court may use extrinsic evidence as an aid in construing the claim language. *Id.* at 1585. The Court may always consider extrinsic evidence from expert witnesses "for background and education on the technology" at issue. *Key Pharmaceuticals v. Hercon Laboratories Corp.*, 161 F.3d. 709, 716 (Fed. Cir. 1998). Further, a trial court may rely on expert testimony in construing a claim when the intrinsic evidence does not answer a particular question. *Id.*; *Vitronics*, 90 F.3d at 1584. A court may only rely on such extrinsic evidence so long as it does not vary on the claim construction mandated by the claims and the patent itself. *Key Pharmaceuticals*, 161 F.3d at 716.

III.

TECHNICAL BACKGROUND RELEVANT TO THE '035 INVENTION

Background information on television tuners is helpful to understanding the inventions claimed in the '035 Patent.[2] This section discusses some basics on radio frequency (RF) signals, the basic concept of a television tuner, how the components of television tuners function, the problems the '035 invention addresses, and some basics on the '035 Patent. Although the '035 Patent is described in the context of a television tuner used to receive analog broadcast television signals, the '035 Patent inventions apply equally to television tuners used for digital television, cable modems, and other digital data systems that utilize standard television channels. *See* Lechner Decl. at ¶¶ 19-21.

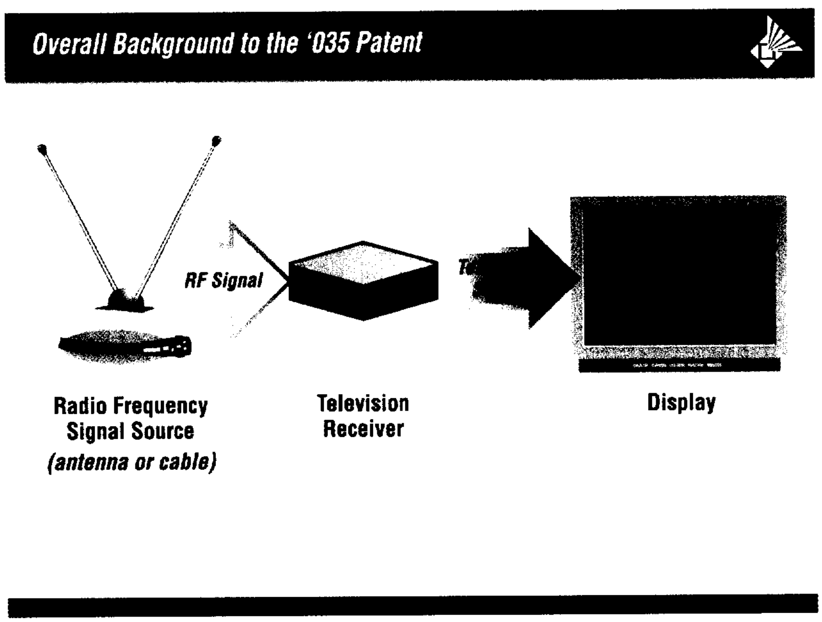
A. Background to the Invention

In the U.S. and abroad, television programs are transmitted by signals having frequencies in the radio frequency (RF) range (discussed below). The RF signals may be sent over-the-air by a transmitting antenna or over a cable in the case of cable television. RF signals carrying

[2] This section summarizes a more detailed technical discussion found in the Declaration of Bernard J. Lechner, filed herewith.

available television channels are input to a television set via a receiving antenna or a cable television system. The television contains a device, called a "television tuner" or "television receiver" which allows the user to extract a single channel from the RF signal. While Graphic Exhibit 1, below, shows a "television receiver" and "display" separately to emphasize the different functions of each, typically both components are within a single television set. *See* Lechner Decl., at ¶ 22.

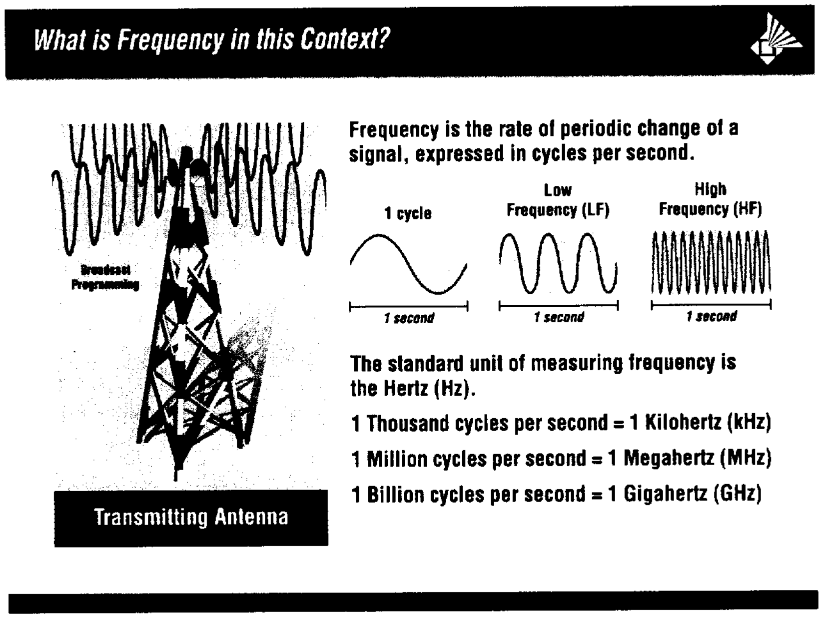
Graphic Exhibit #1



A basic but important concept used throughout the '035 Patent is that of the "frequency" of an electrical signal. "Frequency" is the periodic rate of change of an electrical signal or other periodic event, expressed in cycles per second, or Hertz (Hz). Since the frequencies of interest in the RF range are very high, conventions have been developed to refer to them with alternate units: Kilohertz (abbreviated kHz), Megahertz (abbreviated MHz) and Gigahertz (abbreviated GHz). One Kilohertz is one thousand Hertz; one Megahertz is one million Hertz; and one Gigahertz is one billion Hertz. The term "radio frequency" (RF) refers electrical signals with at frequencies between 100,000 Hz (or 100 kHz) to about 300,000 MHz (or 300 GHz), as shown below in Graphic Exhibit 2. The range of RF frequencies is sometimes referred to as the "RF spectrum" or "RF band of frequencies." Within the RF spectrum, many different types of signals

are present. For example, signals relating to television, FM radio, cellular telephones, pagers and most radar systems all occur between 30 MHz and 3 GHz. *See* Lechner Decl., at ¶¶ 23-27.

Graphic Exhibit #2



For broadcast television, over-the-air television signals in the United States are present between about 54 MHz and 890 MHz.[3] *See* Ex. A at col. 8, ll. 40-51. This range is sometimes called the "broadcast television spectrum." Cable television has a slightly different spectrum, which is currently defined to run from 54 MHz to 864 MHz. Graphic Exhibit 3, below, shows that television occupies only a small part of the RF spectrum. Also shown in Graphic Exhibit 3 is the band of frequencies for FM radio, which occupies much less space than the television band. *See* Lechner Decl., at ¶ 27.

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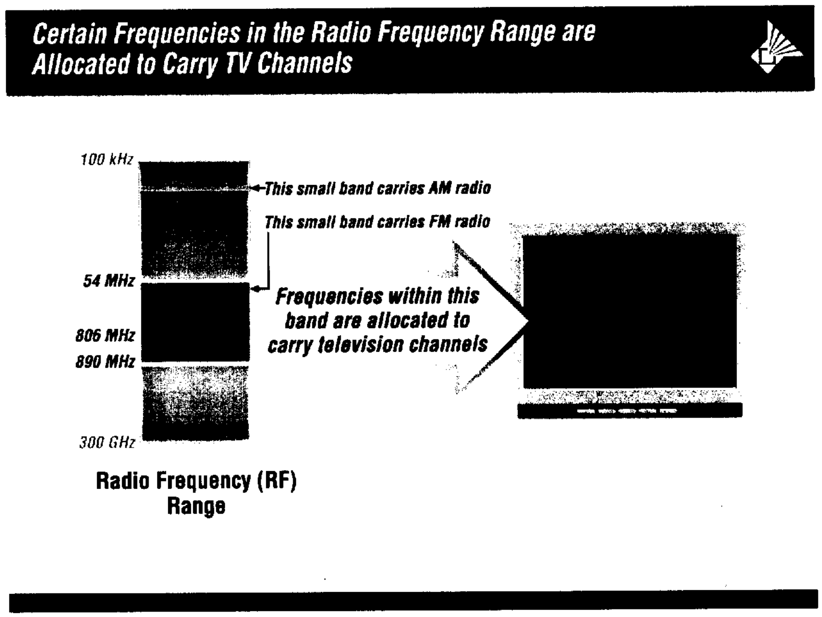
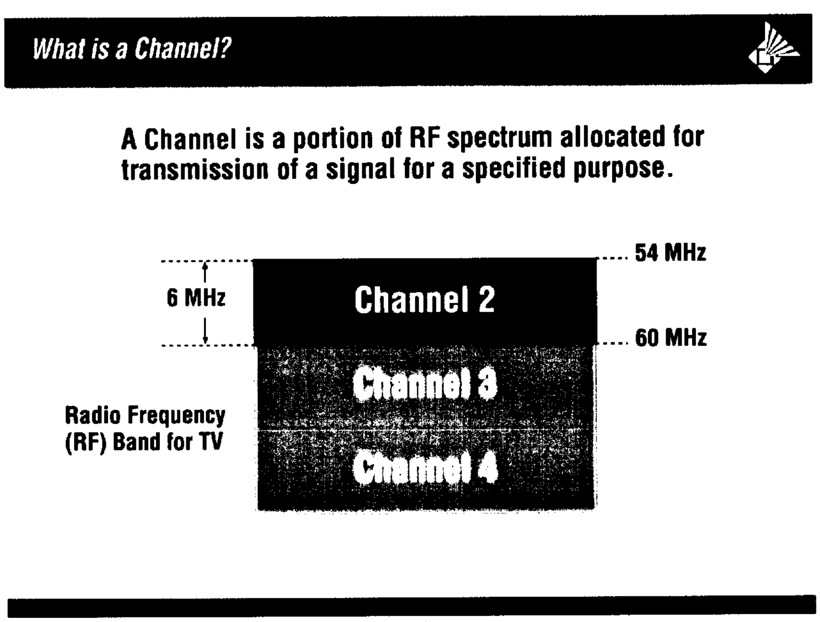
[3] Receiver coverage of channels 70 through 83 has been voluntary since July 1982, when the band from 806 to 890 MHz was reallocated by the FCC. Existing stations in this band were permitted to operate indefinitely. *See* Lechner Decl. at ¶ 27.

Graphic Exhibit #3



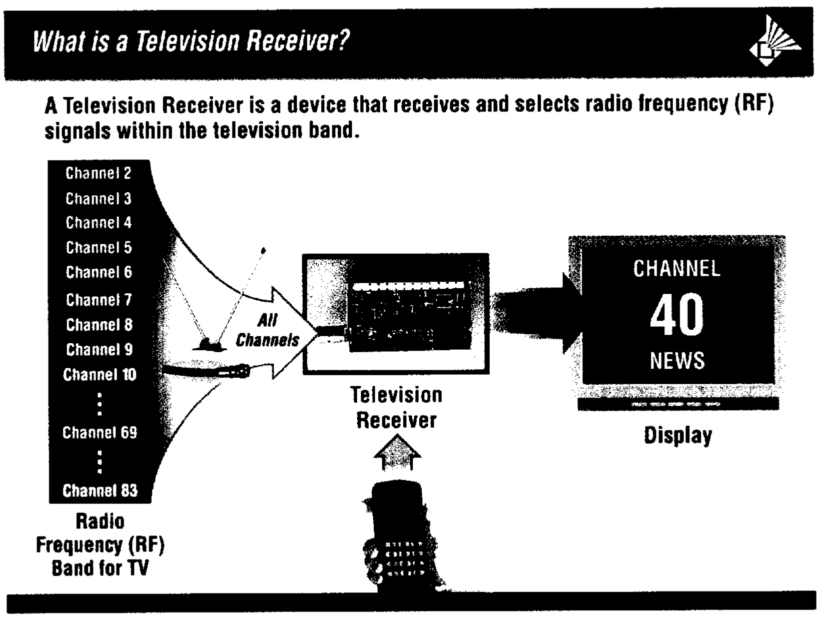
Within the broadcast television spectrum, many different channels are present. Each channel occupies its own band of frequencies. Each television channel in the United States is allocated 6 MHz of bandwidth, as shown in Graphic Exhibit 4, below. *See.* Exhibit A, at col. 8, ll. 31-41. The channel allocations for cable television are slightly different, as the band ranges from 54 MHz to 864 MHz. Within the band used for cable television several channels are used for cable modem data. *See* Lechner Decl., at ¶¶ 28-29; 59-60.

Graphic Exhibit #4



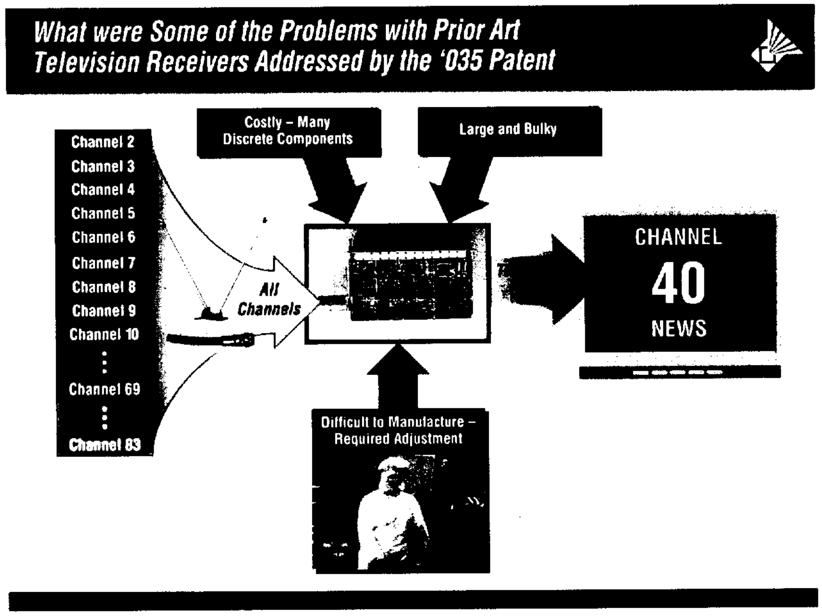
Since a television set receives many television channels (each carried by a band of frequencies), a device is needed to select a channel for viewing or, in the case of a channel carrying only digital data, for processing by a computer. The device which selects and extracts a channel from the television band is called a "television receiver" or "television tuner." Graphic Exhibit 5 illustrates a television receiver (which is typically located inside a television set) being used to select channel 40. *See* Lechner Decl.,. at ¶ 29.

Graphic Exhibit #5



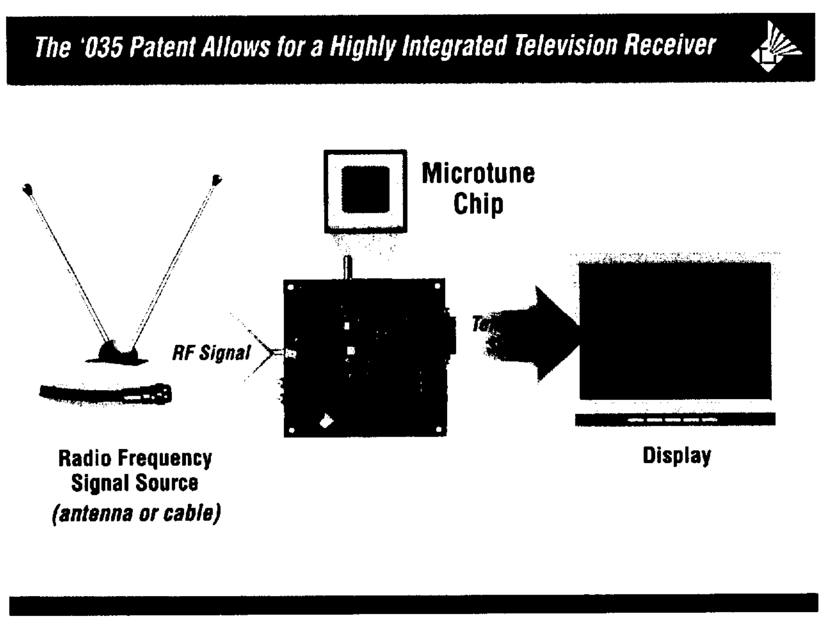
Conventional television tuners contain many individual circuit components, and are difficult to manufacture because these components require manual adjustment as part of the manufacturing process. In addition, a conventional tuner is too large and bulky to use in many electronic devices that now receive television signals, such as laptop computers. The disadvantages of conventional tuners are shown in Graphic Exhibit 6. *See* Lechner Decl., at ¶ 30.

Graphic Exhibit #6



The '035 Patent is directed to making a highly integrated television receiver, i.e., most of the receiver is present on a single integrated circuit, or chip, as shown in Graphic Exhibit 7. A chip-based receiver minimizes the number of individual circuit components, minimizes adjustments, and decreases size and cost. *See* Lechner Decl., at ¶ 31.

Graphic Exhibit #7



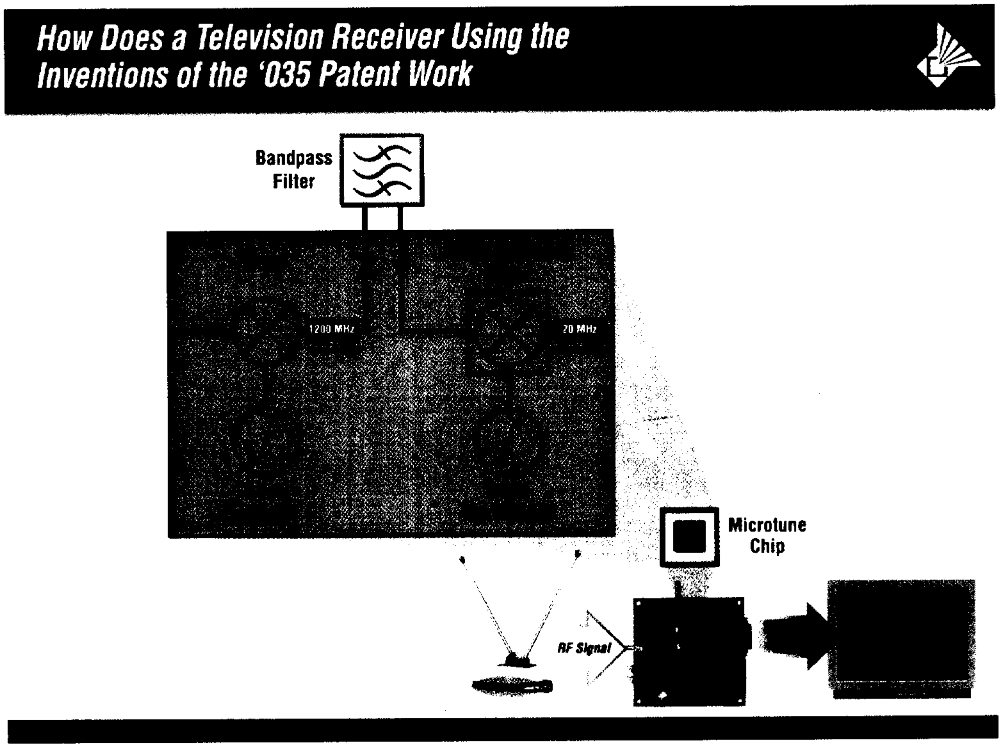
B. The '035 Patent Technology

The '035 Patent concerns using a highly integrated television receiver to receive television signals (from an antenna or cable television system), and then to process the television signal to obtain a single channel for viewing, in the case of a television set. Basically, a television receiver accomplishes the reverse of the television transmission process, in which a television signal is modulated onto a carrier signal at a specific frequency to create a distinct television channel. The function of a television receiver is to select the particular frequency range allocated for the desired channel, convert it to a predetermined intermediate frequency (IF) to facilitate the rejection of adjacent channels and thereby reduce interference, and ultimately to extract the original information that was transmitted, whether that be video and audio signals or data transmitted over a television channel, as in the case of a cable modem. *See* Lechner Decl., at ¶¶ 32-33.

In order for a television receiver to select one channel from a broad range of television channels at its input, a number of components are utilized, each of which are explained below. The components which can be used for this processing are shown schematically in Graphic Exhibit 8, below.

The components shown in Graphic Exhibit 8 are not all claimed in every asserted claim of the '035 Patent. To understand what is claimed, however, this brief will first explain how each component works (mixer, local oscillator, image rejection mixer, filter), and then describe how a signal is processed according to the inventions of the '035 Patent.

Graphic Exhibit #8

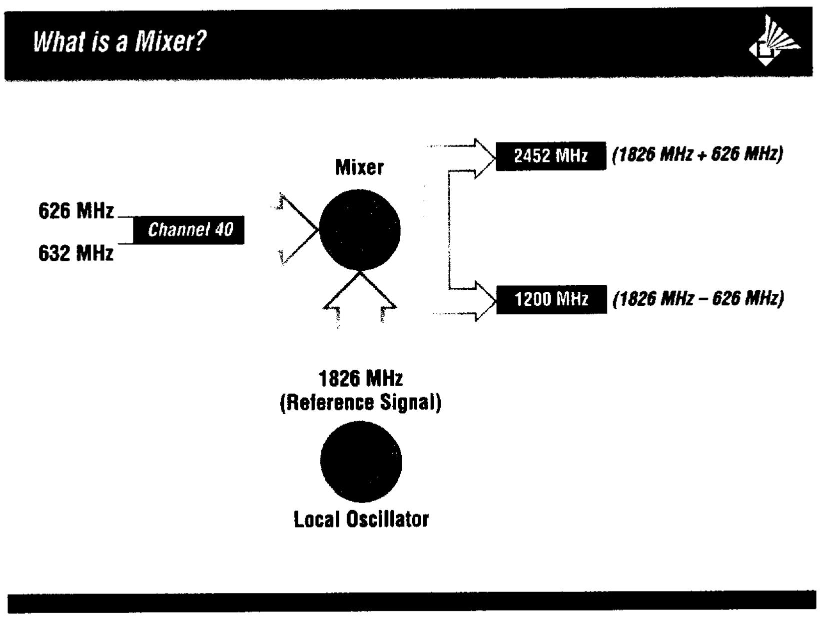


The first component shown is a "mixer." The purpose of the mixer here is to convert the frequency of the selected channel (Channel 40 for example) to a predetermined intermediate frequency (1200 MHz for example) so the selected channel can be separated from the undesired channels and further processed. A "mixer" combines two input signals to create an output signal containing new frequencies. The frequencies contained within the mixer output signal are mathematical products of the frequencies contained within the two mixer input signals. The output signal of a mixer contains components whose frequencies are the sum and difference of the frequencies of the input signals. A mixer output signal also contains other components, including unwanted distortion components, that occur at other frequencies related to the input frequencies. *See* Lechner Decl., at ¶¶ 34-36.

Graphic Exhibit 9, below, shows basically how a mixer works. Note that a "mixer" is denoted by a circle with an "x" in the middle. One input is "626 – 632 MHz" (the frequency band allocated for channel 40) and the second input is "1826 MHz (Reference Signal)." The

1826 MHz signal is shown coming from a "local oscillator" (described below), which is depicted as a circle with a wave inside it. The output of the mixer is the sum of the two input frequencies and also the difference between the two input frequencies. In determining the difference between the two input frequencies, the lower frequency is always subtracted from the higher frequency. *See* Lechner Decl., at ¶¶ 37-39.

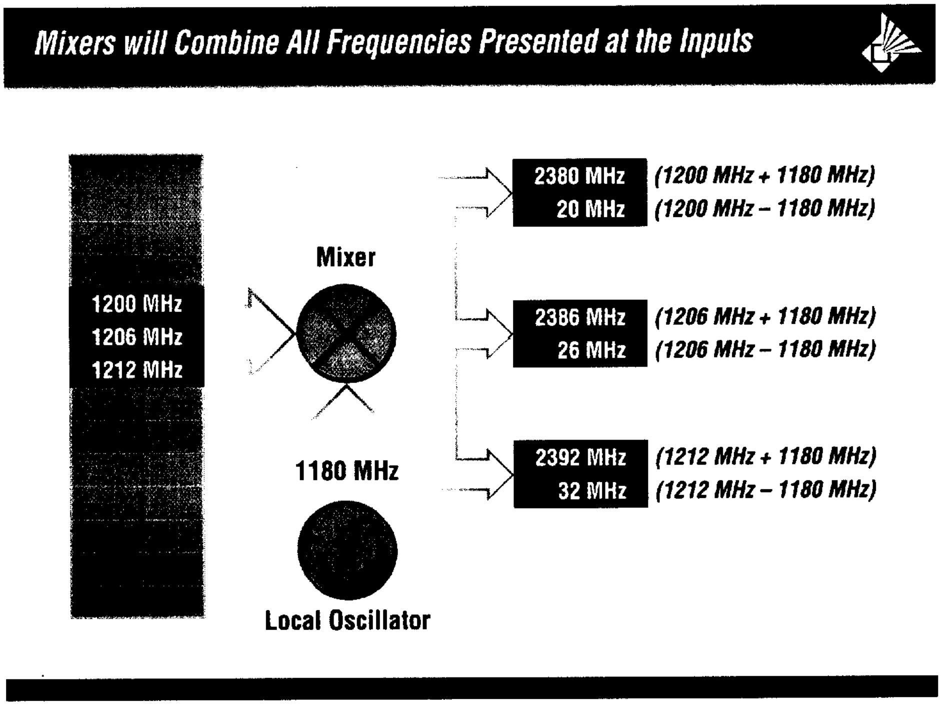
Graphic Exhibit #9



A "local oscillator" is a standard device, well known in the television art, which generates electrical signals of a certain frequency. A local oscillator may be "fixed," in which case it generates a signal at one frequency only, or "variable," in which case it can generate a signal at any frequency within a band of frequencies. By using a local oscillator as one of the inputs to the mixer, the frequencies of the signals present at the output of the mixer can be varied in direct relation to the change in frequency of the local oscillator input signal. *See* Lechner Decl., at ¶ 40.

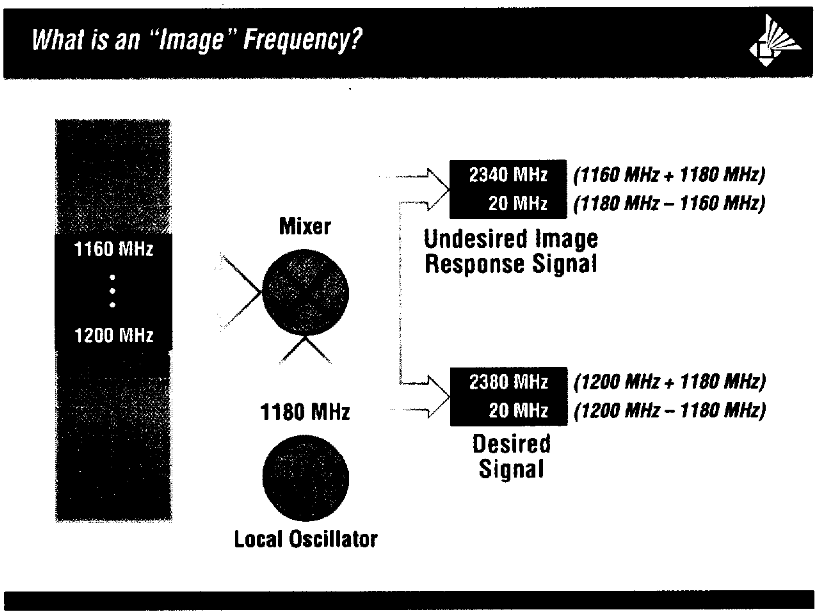
One characteristic of a mixer is that it combines all frequencies presented at its inputs into the output signal according to the rules above. Thus, if a range of frequencies is presented at its input, a mixer will create "ranges" of frequencies at its output, as shown in Graphic Exhibit 10, below. *See* Lechner Decl., at ¶ 41.

Graphic Exhibit #10



Because a mixer operates on a range of input frequencies (for example, television channels 2 through 83, as shown), the mixer can generate an undesired output, called an "image" response. The basic problem is that two different input frequencies, one corresponding to the desired signal and the other corresponding to the "image" or "image signal," can be converted to the same output frequency by the mixer. If a signal exists at the "image" frequency of the selected channel, interference will result. An "image" is shown below in Graphic Exhibit 11. *See* Lechner Decl., at ¶¶ 42-44.

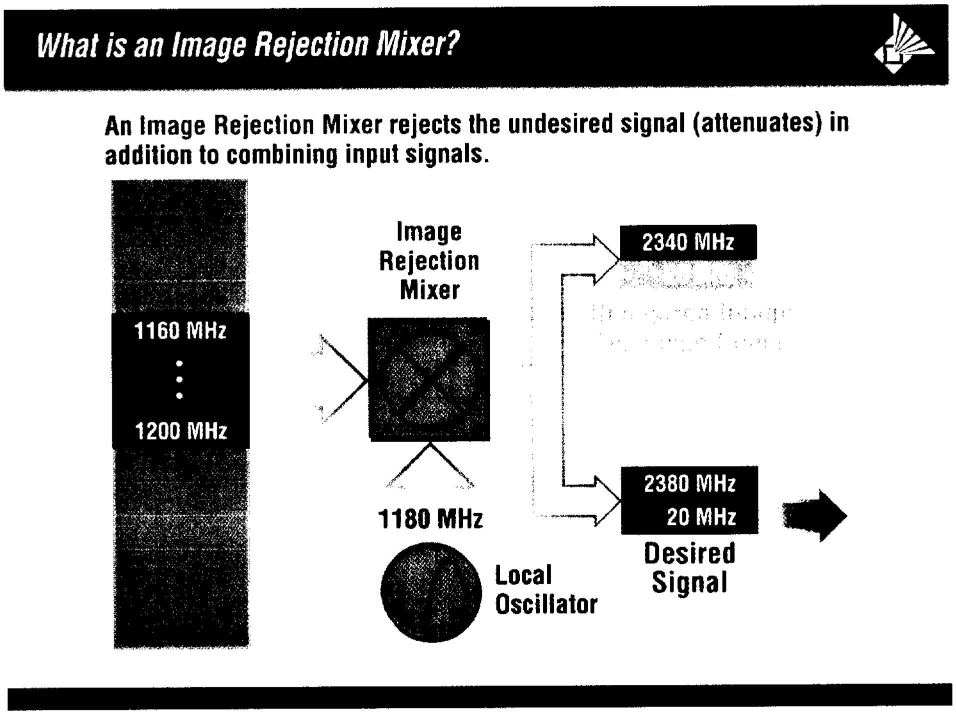
Graphic Exhibit #11



As shown in Graphic Exhibit 12 below, signals spanning a range of frequencies, including the spectrum from 1160 MHz to 1200 MHz, are fed into one mixer input. Assume 1160 MHz corresponds to channel 34 and 1200 MHz corresponds to channel 40, the selected channel. The reference signal is at 1180 MHz. As the selected channel is at 1200 MHz, the mixer will convert the output signal for that channel to the desired second intermediate frequency (IF) at 20 MHz (1200 MHz - 1180 MHz = 20 MHz). However, another frequency present at the input to the mixer is at 1160 MHz (representing signal energy within channel 34 in this example), and will also mix with the 1180 MHz reference signal to create a 20 MHz output from the mixer (1180 MHz – 1160 MHz = 20 MHz). Thus, two channels (the selected and the undesired) will be mixed to the same intermediate frequency, thereby causing interference with the desired signal. *See id.*

A special type of mixer, called an "image rejection mixer" is designed to eliminate or attenuate (reduce) the undesired image signal. An image rejection mixer is shown below in Graphic Exhibit 12.

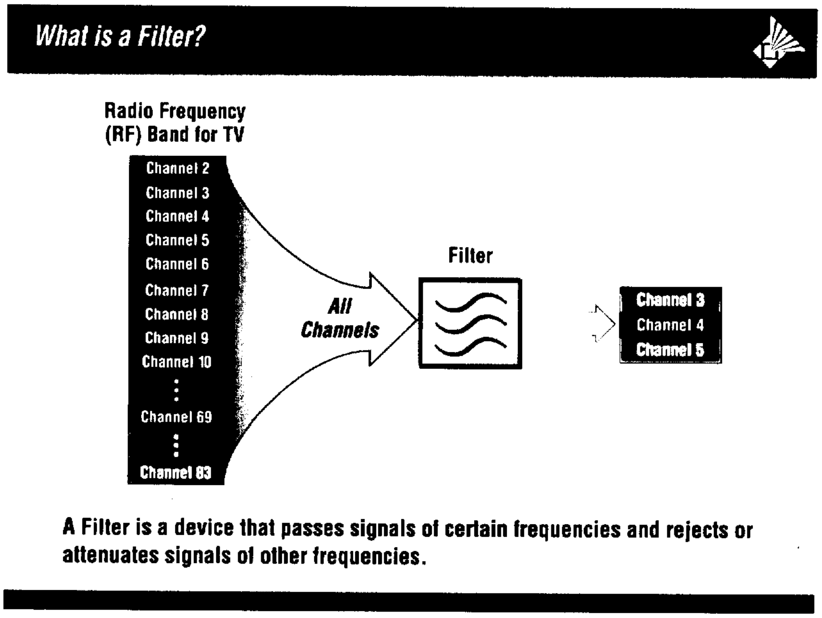
Graphic Exhibit #12



As shown above, an image rejection mixer is symbolized by a circle with an "x" in the middle, located inside a square. This is shown in Figure 4 of the '035 Patent. An image rejection mixer is specially constructed to attenuate the undesired image signal, and thereby prevent or reduce the undesired interaction of the image response with the desired output signal. *See* Lechner Decl., ¶¶ 45-46.

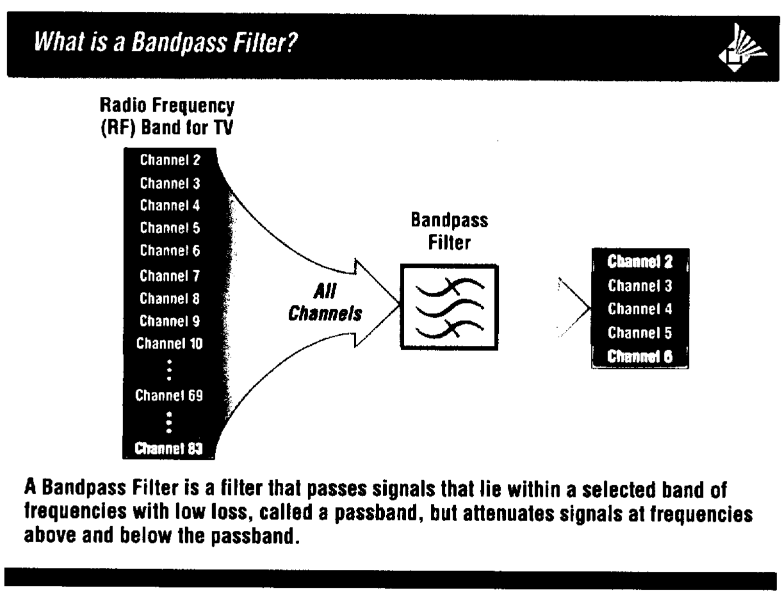
Another component disclosed in the '035 Patent (and known in the industry) is a "filter," as shown below in Graphic Exhibit 13. A filter is designed to pass only signals are certain frequencies and attenuate signals at all other frequencies. For example, a filter could be designed to pass channel 4 and attenuate channels 2-3 and 5-83, as shown. One use for a filter in a television tuner is to eliminate (filter out) the undesired image signal at the input to a mixer. *See* Lechner Decl., at ¶ 47.

Graphic Exhibit #13



There are different types of filters. For example, a "bandpass filter" is a type of filter that passes signals that lie within a selected band of frequencies, called the "passband" and attenuates signals at frequencies above and below the passband. Graphic Exhibit 14 shows a bandpass filter which passes three channels (in this case channels 3-5). A bandpass filter can be used in conjunction with an image rejection mixer to augment the image rejection. *See* Lechner Decl., at ¶¶ 48-49.

Graphic Exhibit #14

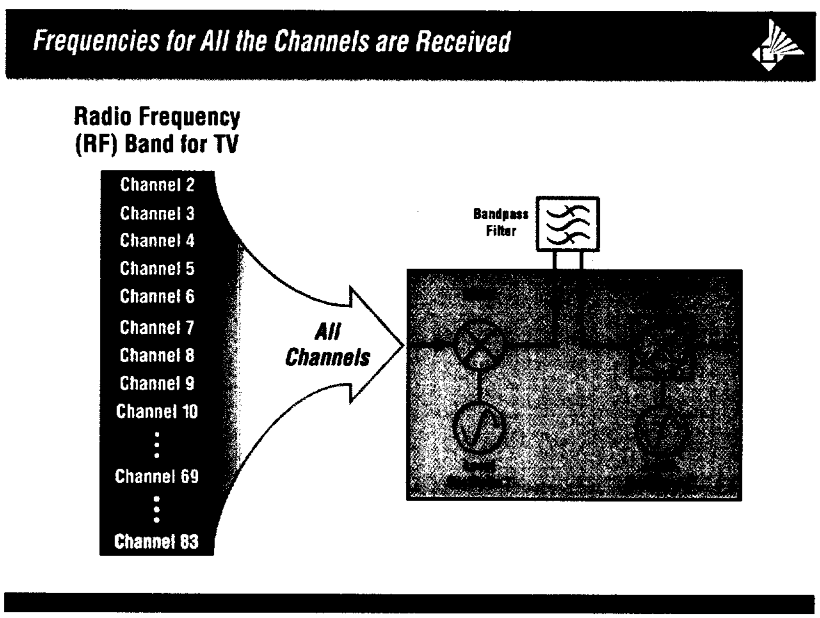


C. Operation of the Invention.

This section discusses how a device built according to the invention in the '035 Patent operates. This discussion is for overview purposes only, and includes elements present in some claims but not all claims. In other words, not all claims require all the components that are described. The example provides an overview of the more difficult concepts of the invention, and traces the path showing how a signal (representing channel 40) is received, mixed, filtered and mixed again.

The input to the television tuner is typically the entire television RF band, namely from 54 to 890 MHz, as shown in Graphic Exhibit 15. In some embodiments of the invention, the RF signal is filtered by a low-pass filter to prevent signals at frequencies greater than 900 MHz from entering the tuner.

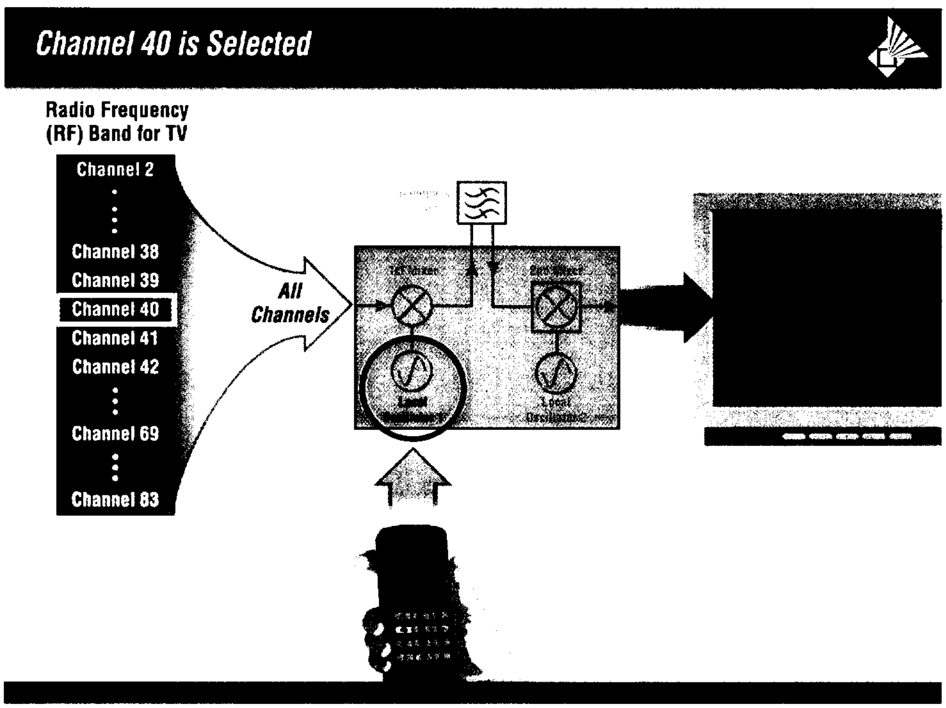
<u>Graphic Exhibit #15</u>



In this example, the viewer desires to view channel 40. The viewer selects channel 40 by using a remote control as shown in Graphic Exhibit 16. When the user selects channel 40 using the remote control shown below, the frequency of the reference signal created by the first local oscillator is set accordingly. The first local oscillator, shown with a red circle, is tuned to cause the frequency band associated with channel 40 to be converted to the desired first mixer output at

1200 MHz, known as the first intermediate frequency (IF). So, as a result of the adjustment to the local oscillator, the frequency associated with channel 40 is converted to 1200 MHz. In this same manner, the selection of any arbitrary channel results in the selected channel being frequency-converted to the same 1200 MHz first IF. It is advantageous to convert all channels to one predetermined frequency, because it permits the use of fixed filters, such as the bandpass filter between the first and second mixers, which can be constructed with greater accuracy and lower cost than filters that must themselves be tuned across some range of frequencies. *See* Lechner Decl., at ¶¶ 50-52.

Graphic Exhibit #16

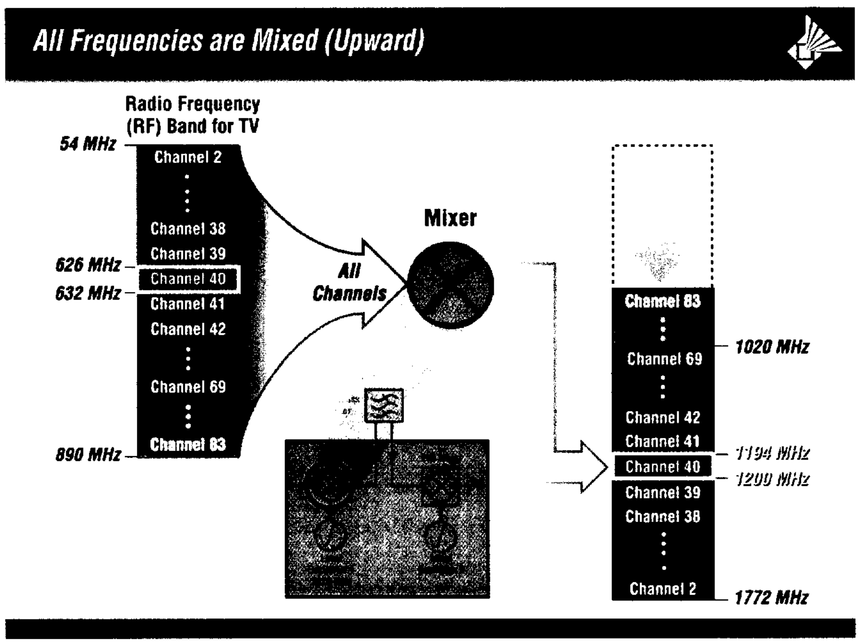


Channel 40 in the United States is located in the UHF band between 626 and 632 MHz. In order to cause this channel to be converted to a first IF of 1200 MHz, it is necessary to tune the first local oscillator to 1826 MHz[4], which will result in the first mixer creating two

[4] As detailed in the Declaration of Bernard Lechner, in reality the actual frequency tuned is 1.25 MHz higher, but for the sake of clarity round numbers are used to illustrate operation of the television receiver.

components: a difference component of 1200 MHz and a sum component of 2452 MHz, the latter not being utilized. A side effect of using the difference component is that the channel is inverted, such that the base of the channel in the RF spectrum (626 MHz) is now the top of the channel at 1200 MHz, and conversely the top of the channel (632 MHz) in the RF spectrum is now converted to 1194 MHz, as shown in Graphic Exhibit 17 below. *See* Lechner Decl., at ¶ 53.

Graphic Exhibit #17

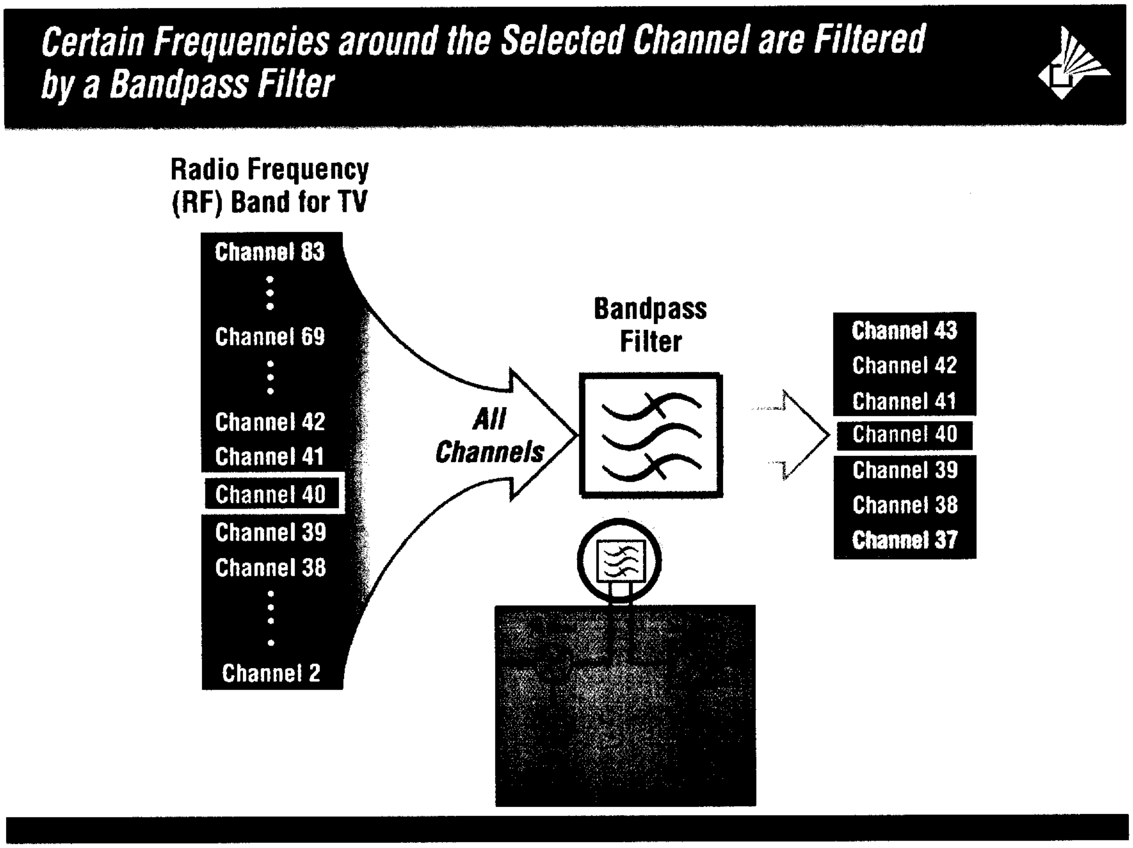


Since, as stated above, the television tuner has as its input the entire television RF spectrum, the output of the first mixer will contain an even broader spectrum of signals at new frequencies. Because signals at frequencies other than the desired channel (channel 40 in this case) will cause interference and distortion in other parts of the tuner, one or more bandpass filters may be used between the first and second mixers to remove as many of these unwanted signals as possible. Note that not all the claims require a bandpass filter.[5] As depicted in Graphic Exhibit 18 below, the bandpass filter between the first and second mixers passes channel

[5] For example, of the claims asserted as infringed, independent claim 120 and claims depending therefrom do not require a bandpass filter.

40, the desired channel, as well as two channels above and below channel 40. All other channels are filtered out, thus preventing interference in other parts of the tuner. *See* Lechner Decl., at ¶ 54.

Graphic Exhibit #18



The output of the bandpass filter, consisting of five channels in this example, is the input to the second mixer, an image rejection mixer, in addition to the second local oscillator reference signal at a frequency of 1180 MHz, as depicted in Graphic Exhibit 19 below. Note that not all the claims require the second mixer to be an image rejection mixer.[6] Unlike the variable frequency first local oscillator, the second local oscillator is fixed at 1180 MHz. The second mixer will create two new signals from the interaction of the desired channel at 1200 MHz and

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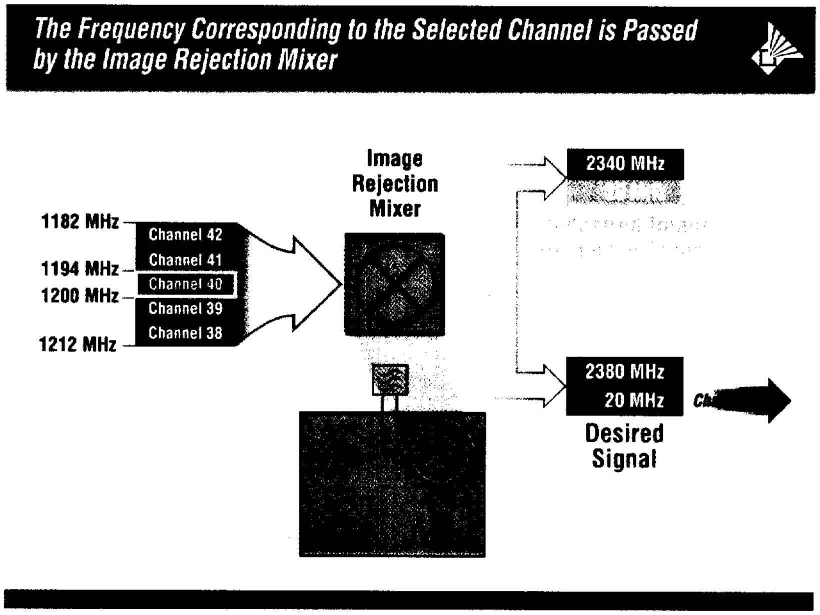
[6] For example, of the claims asserted as infringed, independent claims 102 and 108 and claims depending therefrom do not require an image rejection mixer.

the local oscillator at 1180 MHz; a difference component at 20 MHz, the desired second IF, and a sum component at 2380 MHz, which is not utilized. *See* Lechner Decl., at ¶ 55.

Graphic Exhibit #19

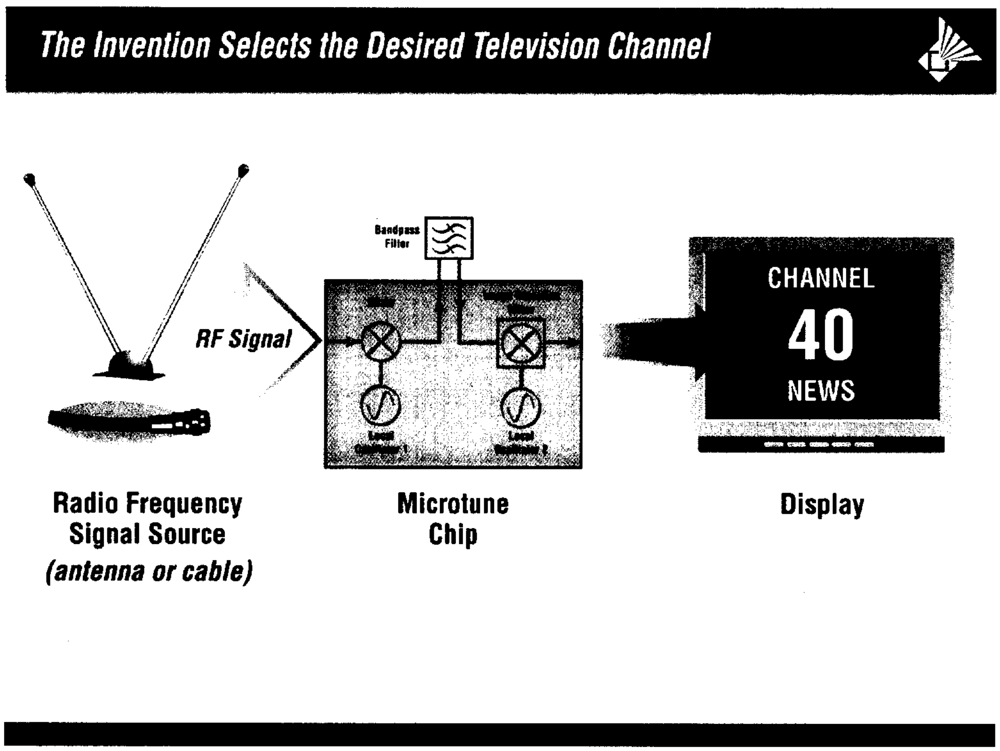


As illustrated in this example, the second mixer is an image rejection mixer, which is required for some but not all of the claims. The purpose of the image rejection mixer is to attenuate an image signal at 1160 MHz, which would be harmful because that signal will combine with the local oscillator at 1180 MHz to produce a mixer output signal that is also at 20 MHz. Note that although this 1160 MHz image falls outside the five-channel passband of the bandpass filter discussed in the example above, in practice some signal energy remains at the image frequency, which is further attenuated by the image rejection mixer. This second 20 MHz signal would cause an undesired signal to be placed on top of the desired signal at 20 MHz and thereby cause interference with the desired signal. However, the image rejection mixer attenuates the image to reduce the interference. If the image were not properly attenuated, it would result in signal degradation, which, depending on the nature of the signal (television program or data), would cause visible picture disruption or data loss. The degree of image rejection achieved by the image rejection mixer is complemented by the image rejection achieved by the bandpass filter. Note that the image rejection mixer still produces the sum

component associated with an input signal present at 1160 MHz, or 2340 MHz, but this is of little concern and easily filtered out of the mixer output. *See* Lechner Decl., at ¶¶ 56-57.

Graphic Exhibit 20 below shows a high-level view of the operation of the television tuner. The large spectrum of television RF signals is selectively tuned until the output consists of only the desired channel, which is then further processed and decoded in subsequent circuitry to extract video and audio baseband signals in the case of analog television broadcasts, or to demodulate data in the case of digital television or cable modems. *See* Lechner Decl., at ¶ 58.

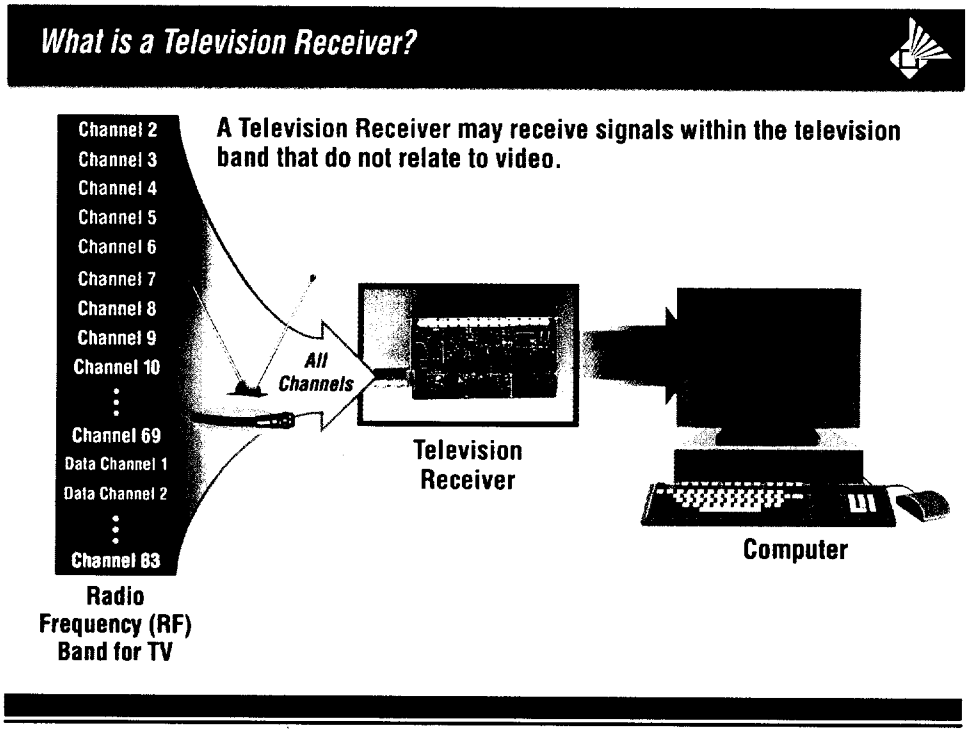
<u>Graphic Exhibit #20</u>



Although the operation of the television tuner inventions of the ‘035 patent have thus far been discussed in the context of a television set receiving video and audio signals, it is important to note that the television tuner disclosed in the ‘035 patent is advantageously used for other applications. For example, Graphic Exhibit 21 below shows the television tuner providing a tuned television signal to a computer for further processing. The television signal in this case would likely be comprised of digital data modulated in a Quadrature Amplitude Modulation (QAM) scheme. Standard 6-MHz television channels have been used to carry data in addition to

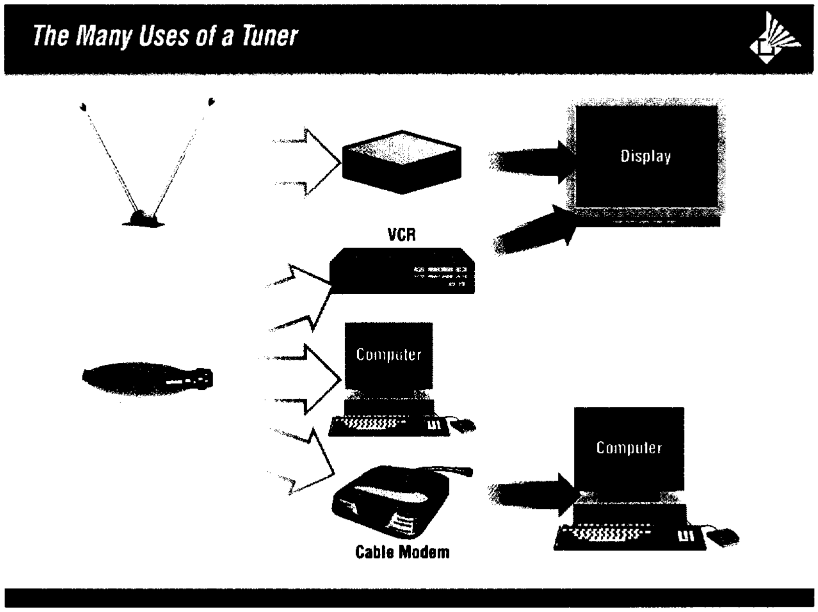
broadcast video/audio signals since the 1970s, and it is becoming commonplace to utilize television channels to carry data exclusively, such as digital cable telephony and two-way Internet connectivity for homes and businesses. *See* Lechner Decl., at ¶ 59.

Graphic Exhibit #21



Various other uses for television tuners are shown in Graphic Exhibit 22 below. In addition to the television set and direct computer connections discussed above, the '035 television tuner would also be advantageously used in video cassette recorders (VCRs) and in cable modems, which convert QAM television signals into packetized digital data for use on computer networks. The scarcity of RF bandwidth available for new purposes assures that additional uses for conventional television signals will be found in the future. What all of these different uses of television signals share in common is the need for highly integrated, high performance, low-cost television tuners as disclosed in the '035 patent. *See* Lechner Decl., at ¶ 60.

Graphic Exhibit #22



IV.

SPECIFIC CLAIM TERMS

This section sets out the key claim terms and their proper construction, using the rules of claim interpretation previously discussed.[7]

A. Television receiver (claims 83-96, 120-128).

Definition.

The term "television receiver" means a device that receives and selects radio frequency (RF) signals within the television band, and is synonymous with "television tuner."

Support for Definition.

This term appears in the preamble of claims 83 and 120, which recite "A television receiver comprising . . . ." It is well-settled that "[i]f the claim preamble, when read in the context of the entire claim, recites limitations of the claim, or, if the claim preamble is "necessary to give life, meaning, and vitality" to the claim, the claim preamble should be

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[7] The parties stipulated to the construction of many claim terms as set forth in Exhibit C. *See* Declaration of John E. Giust in Support of Microtune (Texas) L.P.'s Markman Brief, Exhibit 1.

construed as if in the balance of the claim." *Pitney Bowes, Inc. v. Hewlett-Packard Co.*, 182 F.3d 1298, 1305 (Fed. Cir. 1999), *quoting Kropa v. Robie*, 187 F.2d 150, 152 (CCPA 1951).

Here, the term "television receiver" gives life, meaning and vitality to the claim because it illuminates what is meant by the express claim limitations. For example, a "channel" (claim 83) refers to a television channel, and an "RF signal source" (claims 83, 120) refers to a source of television signals. Therefore, the preamble further identifies and gives meaning to the subsequent claim elements, which means that the preamble is a limitation of claims 83 and 120.

In the prosecution history, and in the patent and its claims, the phrases "television tuner" and "television receiver" are used interchangeably. In fact, during prosecution of the '035 Patent, the Examiner characterized U.S. Patent 4,162,452 to Ash (the "Ash reference") as a "television receiver." Lechner Decl., Ex. 3, at 5. This characterization was echoed by the applicant. Lechner Decl., Ex. 4, at 11. Strictly speaking, the Ash reference discloses a television tuner, making it apparent that the term "television receiver" should be construed in the same manner as a "television tuner."

The term "television tuner" is frequently referred to within the specification, indeed over fifty times. *See, e.g.,* Ex. A at col. 1, l. 6, col. 2, l. 39, col. 2, l. 58. Moreover, the very title of the '035 Patent is "Highly Integrated Television Tunr on a Single Microcircuit" and the technical field of the invention is described as "television tuner circuits." *Id.* at col. 1, l. 6. However, the claims use the phrase "television receiver" in the same manner, and should be given the same definition.

A "television tuner" is known in the art as a device that receives and selects radio frequency (RF) signals within the television band, and this is the definition that should be given to "television receiver." *See* Lechner Decl. at ¶ 63.

1. Channel (claim 83).

As stated above, the recitation of "television receiver" in the preamble of claim 83 gives meaning to the term "channel" in claim 83. Therefore "channel" in claim 83 means a "television" channel, more specifically "a band of frequencies allocated for carrying a television

signal, for example a 6 MHz band of frequencies between 54-890 MHz." This is consistent with the patent, Exhibit A, at col. 8, ll. 38-51.

2. Channel (claims 102, 108).

Since the preamble of claims 102 and 108 do not indicate a television device operating on television frequencies, "channel" in these claims is simply "a portion of RF spectrum allocated for transmission of a signal for a specified purpose" as stated in Graphics Exhibit 4, *supra*.

3. RF signal (claims 83, 120).

As stated above, the recitation of "television receiver" in the preamble of claims 83 and 120 gives meaning to the term "RF signal" source in claim 120 as a source of television RF signals. Therefore "RF signal" in claim 120 means "signals allocated for television, for example, from 54-890 MHz". This is consistent with the patent, Exhibit A, at col. 8, ll. 38-51.

B. Image, Image Signal (claims 124 and 125).

Definition.

The terms "image" and "image signal" are synonymous and mean an undesired signal, that, along with the desired signal, is present at the signal input of a mixer. The image signal is located at a frequency that is twice the frequency of the desired mixer output signal, namely the intermediate frequency (IF), away from the frequency of the desired signal. The result, in the case of a basic mixer, is to create a second output signal at the same frequency as the desired output signal, thereby causing interference with the desired mixer output signal.

Support for Definition.

The frequency of an image signal for a given desired mixer input signal frequency and local oscillator input signal frequency can be determined by using a formula derived from the definition above. When the desired mixer output signal is the difference component, if the local oscillator frequency is higher than the frequency of the desired signal, the frequency of the image signal will be higher than the frequency of the desired signal; if the local oscillator frequency is lower than the frequency of the desired signal, the frequency of the image signal will be lower than the frequency of the desired signal. The specification provides ample support for these

terms. For example, it is stated that "[t]he image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF}=|1160-1180|=20 MHz." Ex. A at col. 10, ll. 4-7. Note that application of the formula f_{IM}=(2 x f_{LO})- f_{RF}, where f_{IM} is the image signal frequency, f_{LO} is the local oscillator frequency and f_{RF} is the desired input frequency, yields f_{IM} =(2 x 1180) – 1200 = 1160 MHz, the image signal frequency identified in the specification. The term "image" is used synonymously and consistently with "image signal" in the specification, as in "the image falling on 1160 MHz." *Id.* at col. 10, l. 30. Similarly, during prosecution the applicant used the term "image signal" in reference to the 1160 MHz image to the examiner that "MIX2 [the image rejection mixer] must reject these image signals associated with the first IF signal." Lechner Decl., Ex. 5, at 9. Thus, the terms "image" and "image signal" are properly defined in the specification and used consistently throughout the specification and claims. *See* Lechner Decl. at ¶ 64.

C. Image Rejection, Rejects an Image Signal (claims 83, 124 and 125).

Definition.

The term "image rejection" and the phrase "rejects an image signal" refers to attenuation of an image signal to such a degree as to minimize its undesired interaction, and thereby reduce interference with the desired output signal.

Support for Definition.

The patent specification discloses that "[t]he image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF} =|1160-1180|=20 MHz." Ex. A at col. 10, ll. 4-7. In the example given, the image signal is located at 1160 MHz. *See id.* at col. 9, l. 62 – col. 10, l. 3. In discussing the operating parameters for this image rejection mixer, the specification mentions "50 dB image rejection" which demonstrates significant, but not total, attenuation of the image signal. *See id.* at col. 13, ll. 4-5.

The concept of "image rejection," namely that an image signal is rejected or significantly attenuated to prevent interference with a desired signal, is adequately described in the specification, and that the definition is consistent with image rejection, as it is known in the art.

There are two common methods for performing image rejection, which can be used individually or together for cumulative effect. The first is by filtering, typically with a bandpass filter, such that the image signal falls somewhere outside the passband of the filter. The second method is to use an image rejection mixer, which is a mixer that is specifically designed to pass the desired frequency while attenuating the image signal. When both an image rejection mixer and a filter are used, the total image rejection achieved (usually expressed in decibels, or dB) is the sum of the image rejection performed by each.

The '035 patent discloses three methods of performing image rejection: a bandpass filter, an image rejection mixer, or a combination of a bandpass filter and image rejection mixer. Figure 4 shows using an image rejection mixer alone. The description of Figure 4 discloses that a filter could be used in conjunction with the image rejection mixer. *See id.* at col. 12, ll. 44-46 ("The design will include a bond wire parallel LC network option to further band-limit any image energy at the output of this mixer if necessary.") In context, this discussion suggests to one of ordinary skill in the art that a "parallel LC network" is placed at the output of the first mixer to prevent energy at the image frequency from reaching the input of the second mixer. A "parallel LC network" consists of an inductor and capacitor in parallel and forms a bandpass filter, as known in the art. A "bond wire" is a wire attached to an integrated circuit (IC) that acts as the inductor in the bandpass filter.

As with the "bond wire" image rejection filter, Figure 5 shows a filter 509 between two mixers. The symbol for the second mixer is general and does not specify an image rejection type of mixer. However, for improved image rejection as discussed with respect to Figure 4 (above) the second mixer in Figure 5 could be an image rejection mixer. *See* Lechner Decl. at ¶ 65.

D. Image Rejection Mixer, Image-Rejection Mixer (claims 83 and 120).

<u>Definition.</u>

The terms "image rejection mixer" and "image-rejection mixer" are synonymous and refer to a certain type of mixer which performs image rejection in addition to combining input signals. In other words, an image rejection mixer not only performs the functions of a basic

mixer, but also specifically attenuates the image signal to limit interference with the desired mixer output signal.

Support for Definition.

The '035 Patent discusses the operation of an image rejection mixer in similar fashion to a basic mixer, as "[t]he IF frequencies, given by $f_{IF} = |f_{RF} \pm f_{LO}|$, have carriers at f_{IF}=1200-1180=20 MHZ and at f_{IF}=1200+1180=2380 MHZ." *See.* Exhibit A, at col. 10, ll. 1-3. The specification further states that "[t]he image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF} =|1160-1180|=20 MHz." *Id.* at col. 10, ll. 4-7. A specific implementation of an image rejection mixer is illustrated in Figure 6, the operation of which is described in the specification at column 13, lines 1-14.

In the file history, the applicant discussed an image rejection mixer as a mixer that prevents "extraneous frequencies from interfering with the desired 20 MHz signal." Lechner Decl., Ex. 5, at 9.

The '035 patent specification as originally filed contains both sufficient definition of the necessity for and function of an image rejection mixer, and further provides an example of how to construct an image rejection mixer that would function appropriately to perform image rejection. One example of an image rejection mixer is provided in Figure 6, and its operation described in the specification at column 13, lines 1-14. The image rejection mixer disclosed in Figure 6 is just one method of constructing an image rejection mixer, and there are many other image rejection mixer designs that would necessarily fall within the definition of image rejection mixer as used in the '035 patent. *See* Lechner Decl. at ¶ 66.

E. Subtractive Mixer, Subtractive Mixing, Subtractively Mixing, Subtractively Mixes (claims 84, 87, 105, 108 and 123).

Definition.

A "subtractive mixer" is a mixer in which only the frequency difference signal components created by the combination of the two mixer input signals are present in the output signal, the frequency sum components being removed by a filter. "Subtractive mixing,"

"subtractively mixing," and "subtractively mixes" means combining two mixer input signals in a mixer such that only the frequency difference signal components created by the mixer are present in the output signal, the frequency sum components being removed by a filter.

Support for Definition.

The term "subtractive mixer" is discussed in various places in the specification. For example, "[m]ixer 408 is a subtractive mixer that will subtract the frequency of local oscillator 450 from the input RF spectrum." Ex. A at col. 9, ll. 21-22. The specification illustrates the operation of a subtractive mixer by pointing out that a basic mixer with a 400 MHz signal at its RF input and a local oscillator input of 1600 MHz will create two mixer outputs, one at 2000 MHz, and one at 1200 MHz. *See id.* at col. 9, ll. 23-28. The essence of the "subtractive mixer" is illustrated by the statement that "[s]ince mixer 408 is a subtractive mixers [sic] the 2000 MHz signal will be filtered out and only the 1200 MHz signal will be passed on to mixer 410." *Id.* at col. 9, ll. 30-32.

The phrases "subtractive mixing," "subtractively mixing" and "subtractively mixes" should be construed consistently with the operation of combining two inputs in a "subtractive mixer," as defined above. *See* Lechner Decl. at ¶ 67.

F. Filter, Filters, Filtered, Filtering (claims 83, 91, 93, 102-103, 107-108, 110, 113-115, 120 and 123).

Definition.

A "filter" is defined as a device that passes signals of certain frequencies and attenuates signals of other frequencies. "Filters," "filtered" and "filtering" means processing a signal such that signals of certain frequencies are passed and signals of other frequencies are attenuated.

Support for Definition.

The specification, in describing the operation of a prior art television tuner, states that "filter 304 limits the signals entering television tuner 300 so that a minimum number of undesired signals exist in television tuner 300." *Id.* at col. 5, ll. 50-52. The specification further relates that "[f]ilter 304 also attenuates signals not in a fairly narrow (100 MHz) range about the desired signal. Finally, known interference signals, such as FM broadcast, shortwave service

signals, signals in the intermediate frequency band and Citizen Band radio signals, are specifically rejected by filter 304." *Id.* at col. 5, ll. 54-59. Thus, the term "filter," as used in the claims and specification of the '035 patent, is completely consistent with the concept and definition of a filter as used in the art. Moreover, the verb forms "filters," "filtered" and "filtering" as used in various claims are consistent with the definition of "filter" and are therefore fully supported by the specification. *See* Lechner Decl., at ¶ 68.

G. Bandpass Filter, Passband (claims 83, 93, 102, 107-108, 111-115).

Definition.

A "bandpass filter" is a filter that passes signals that lie within a selected band of frequencies with low loss, called the "passband," but attenuates signals at frequencies above and below the "passband."

Support for Definition.

The specification states, in describing the operation of bandpass filter 304, that "[f]ilter 304 also attenuates signals not in a fairly narrow (100 MHz) range about the desired signal." *Id.* at col. 5, ll. 54-55. In this example, the "passband" is 100 MHz wide, and the bandpass filter attenuates frequencies both below and above this passband. The specification also uses "passband" in describing the upper and lower frequency limits of the passband of Second IF filter 420, which is of the bandpass type. *See id.* at col. 13, ll. 51-53. The terms "bandpass filter" and "passband" are properly defined and used in the specification. *See* Lechner Decl., at ¶ 69.

H. Receiver input "coupled to an RF signal source" (claim 83).

Definition.

The phrase "coupled to an RF signal source" means that an electrical connection is made to a source of one or more RF signals, for example, an antenna receiving over-the-air signals or a coaxial cable for connection to a cable television system. The term "RF signal source" is not an element.

Support for Definition.

By its plain language, the claims require only that the accused product have an input "coupled to" an RF signal source, and the claim does not require such a source to be present. Thus, the phrase is not a claim element. In the specification, operation of the prior art embodiment is described as "[t]elevision tuner 300 receives a standard television RF signal from either antenna 302 or a cable system connection (not shown)." *See.* Ex. A, at col. 5, ll. 47-49. In illustrating the invention of the '035 Patent, the specification states that "[t]he RF signal enters television tuner 100 from the antenna 402 (or cable, not shown) and is passed through RF low-pass filter (RFLPF) 404." *Id.* at col. 8, ll. 32-34. Thus, the specification provides two examples of ways of coupling the television tuner to an RF signal source: an antenna or connection to cable television system. *See* Lechner Decl., at ¶ 70. The accused product need only have an input coupled to such a source.

I. Passing more than one channel (claim 83).

Definition.

The phrase "passing more than one channel" means that the passband of the bandpass filter processing a television RF signal would be greater than one channel, i.e., greater than the 6-MHz bandwidth of a standard television channel in the United States.

Support for Definition.

The specification states "each channel in the VHF and UHF band has a 6 MHz bandwidth." *See.* Ex. A, at col. 8, ll. 43-44. "Passing more than one channel" must necessarily mean passing more than 6 MHz. *See* Lechner Decl. at 34-35. Moreover, during the prosecution of the '035 Patent, the applicant distinguished this limitation of claim 84 (which issued as claim 83) by pointing out that it was patentably distinct over several prior art references, including the Ash reference and U.S. Patent 4,340,975 to Onishi *et al.* (the "Onishi reference"), both of which employed bandpass filters with a passband of one channel or less. *See* Lechner Decl., Ex. 6, at 8; Ex. 4 at 8-9. The Onishi reference discloses a passband of 5 MHz, and the Ash reference discloses a passband of 6 MHz. Lechner Decl., Ex. 11, at col. 5, ll. 65-68; Ex. 12, at col. 10,

ll. 55-59. Therefore, the phrase "passing more than one channel" is adequately supported by the specification. *See* Lechner Decl., at ¶ 71.

J. Having at least four channels (claims 102 and 108).

Definition.

The phrase "having at least four channels" means that the passband of a bandpass filter processing a television RF signal would be greater than four television channels, i.e., 24 MHz, the bandwidth of four contiguous standard television channels in the United States.

Support for Definition.

The phrase "having at least four channels" must necessarily mean passing more than 4 television channels. For example, the specification states "each channel in the VHF and UHF band has a 6 MHz bandwidth." Ex. A at col. 8, ll. 43-44. Four times 6 MHz is 24 MHz, and therefore "having at least four channels" must mean that the bandwidth passed must be equal to or greater than 24 MHz. *See* Lechner Decl. at 35. Moreover, during the prosecution of the '035 Patent, the applicant distinguished this limitation of claim 84 (which issued as claim 83) by pointing out that it was patentably distinct over several prior art references, including the Ash and Onishi references, both of which employed bandpass filters with a passband of one channel or less (Onishi - 5 MHz, Ash - 6 MHz). *See* Lechner Decl., at ¶ 72; Lechner Ex. 6, at 8; Ex. 4, at 8-9; Ex. 11, at col. 5, ll. 65-68; Ex. 12, at col. 10, ll. 55-59.

K. A first bandpass filter...performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent circuitry (claim 83).

Definition.

The phrase "a first bandpass filter...performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent circuitry" means that the bandpass filter accomplishes some detectable measure of image rejection, but less than all of the image rejection accomplished by the invention, and further that the bandpass filter decreases the total power content of the signal it processes, thus reducing the power-handling requirements of circuitry that follows the bandpass filter.

Support for Definition.

This phrase was added to claim 84 (which issued as claim 83) through amendment during prosecution of the '035 Patent to distinguish the invention over prior art television tuners disclosed in the Ash and Onishi references, both of which employed bandpass filters that accomplished **all** image rejection. *See* Lechner Decl., Ex. 6, at 4; Ex. 11 at col. 7, l. 67 – col. 8, l. 4; Ex. 12 at col. 10, ll. 55-59. Filters commonly used between the first and second mixers of television tuners for the purpose of rejecting the image frequency at the input to the second mixer are typically designed to have a narrow passband, such as one channel or 6 MHz. A narrow passband filter usually has greater attenuation within its passband than does a wider passband filter. *See* Lechner Decl. at 36. The bandpass filter recited in claim 83 of the '035 patent performs *some* of the image rejection; with the balance performed by the second mixer, which is an image rejection mixer. Accordingly, the definition of this phrase is consistent with the expected behavior of a bandpass filter used in conjunction with an image rejection mixer, as recited in claim 83. See Lechner Decl., at ¶ 73.

L. "Coupled to" (claim 120).

Definition.

The phrase "a second mixer having a first input coupled to an output of said first mixer" means that the signal path extends from the first mixer output to the second mixer output, and there can be intermediate circuit components between the first mixer output and second mixer input.

Support for Definition.

The plain meaning of "coupled to" means "connected to." However, an electrical component can be connected to another electrical component even if intermediate components are connected in between. *See* Lechner Decl. at 37. In fact, the specification shows a first mixer 408 connected to an intermediate component (a filter 509) and then, connected to a second mixer 510. *See* Figure 5. The specification, in Figure 4, also shows a "direct" connection between mixers 408 and 510 without a filter between the mixers.

The "coupled to" language was present in originally filed claim 1. In its first office action, the U.S. Patent and Trademark Office ("PTO") construed and rejected claim 1. Claim 1 was rejected over the Ash reference, and construed to permit a component (a filter in the case of Ash) to exist between the first mixer output and the second mixer input. *See* Lechner Decl., Ex. 3, at 2 (rejection). In response, the applicant amended claim 1 to recite "coupled directly to." *See* Lechner Decl., Ex. 4, at 2 (insertion of "directly" into claim 1); *Id.* at 7-8 (discussion of addition of word "directly"). This amendment was made to overcome the PTO's construction of claim 1 as permitting intermediate components.

When adding claim 121, now claim 120, the applicant did not use "coupled directly to" but instead kept the original broader wording of "coupled to." Claim 120 was filed during the prosecution of the patent (as claim 121 at that time) after the examiner had already construed "coupled to" broadly. *See* Lechner Decl., Ex. 7, at 2 (adding claim 121, now renumbered as claim 120). The applicant argued that claim 120 was patentable due to the limitations present in allowed claims 10 and 28, which were also inserted into claim 120. Those limitations concerned an "input filter" and an image rejection mixer. *See id.* at 4:

> New independent claim 121 [now claim 120] includes both the limitations of an input filter and an image rejection mixer. Applicant submits that the prior art does not teach or suggest the combination of elements in claim 121, and therefore, claim 121 is patently novel and nonobvious.

Based on the claim language, specification and file history, the claim phrase "coupled to" in claim 120 should be construed broadly to permit the inclusion of other circuit components between the two mixers, such as a filter. *See* Lechner Decl., at ¶ 74.

V.

CONCLUSION

Microtune has construed the claim terms based upon the intrinsic record. Microtune has offered extrinsic evidence only to provide a general technical background of the invention. Exhibit B hereto is a table cross-referencing the claims to Microtune's proposed definitions. Microtune respectfully requests the court to adopt Microtune's definitions as recited in Exhibit B.

Dated: January 22, 2002

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CERTIFICATE OF SERVICE

This will certify that a copy of the foregoing document was served on the following counsel for Defendant by Federal Express, on the 22nd day of January, 2002.

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Exh A

US005737035A

United States Patent [19]

Rotzoll

[11] Patent Number: 5,737,035

[45] Date of Patent: Apr. 7, 1998

[54] HIGHLY INTEGRATED TELEVISION TUNER ON A SINGLE MICROCIRCUIT

[75] Inventor: Robert Rudolf Rotzoll, Allen, Tex.

[73] Assignee: Microtune, Inc., Plano, Tex.

[21] Appl. No.: 426,080

[22] Filed: Apr. 21, 1995

[51] Int. Cl.^{6} H04N 5/44

[52] U.S. Cl. 348/725; 348/731; 455/315; 455/339

[58] Field of Search 455/302, 315, 455/339, 179.1; 348/725, 726, 731; H04N 5/44

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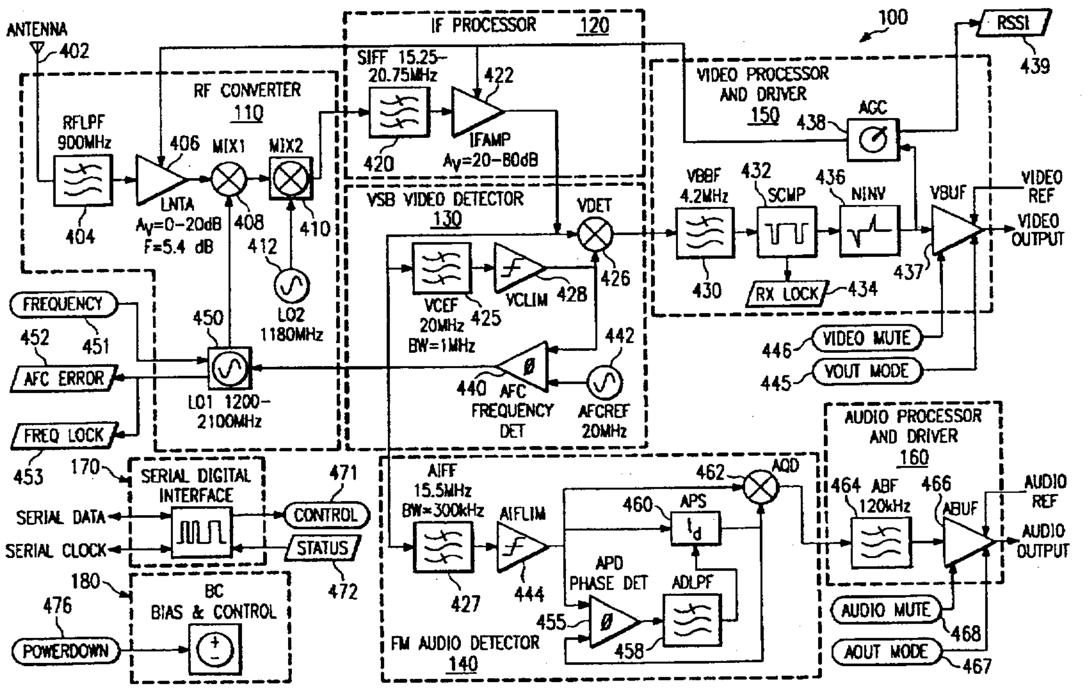
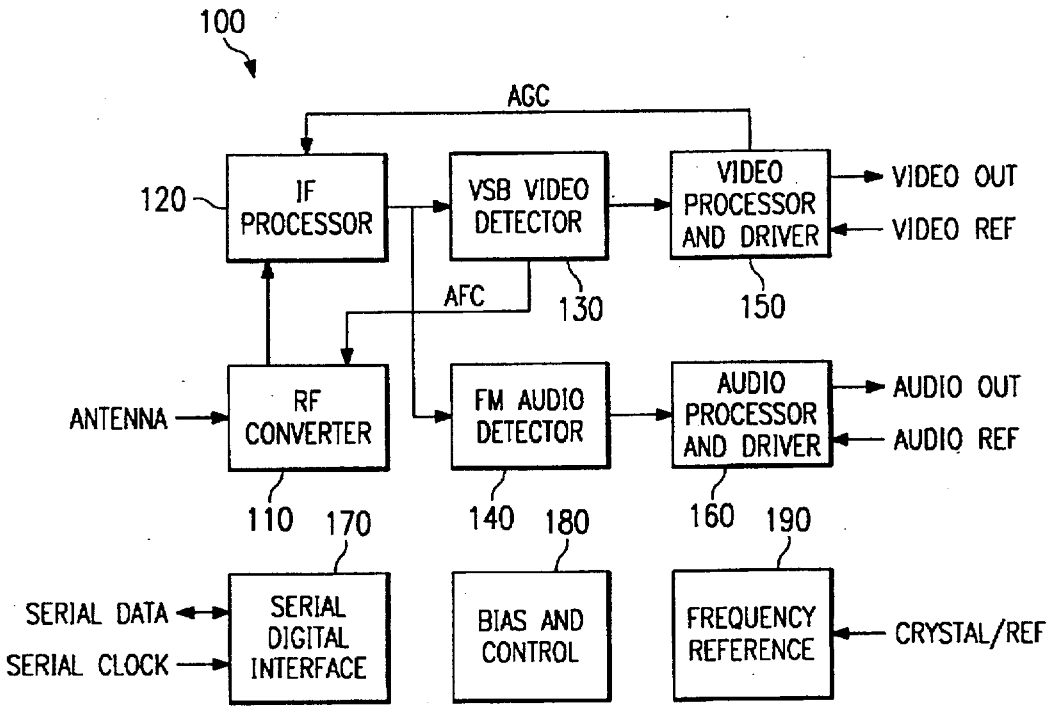
Primary Examiner—Sherrie Hsia
Attorney, Agent, or Firm—Fulbright & Jaworski L.L.P.

[57] ABSTRACT

There is disclosed a fully integrated television receiver for receiving a standard antenna or cable input and outputting a standard video baseband signal and a standard audio baseband signal. The receiver employs an up-conversion mixer and a down-conversion image-rejection mixer in series to produce an on-chip IF signal. Audio detection of the IF signal is performed via an audio delay-locked loop operating in quadrature to the audio FM signal. Video detection of the IF signal is performed by frequency mixing the IF signal with the extracted video carrier of the IF signal. Variable load drive capability is provided for both the video output and the audio output.

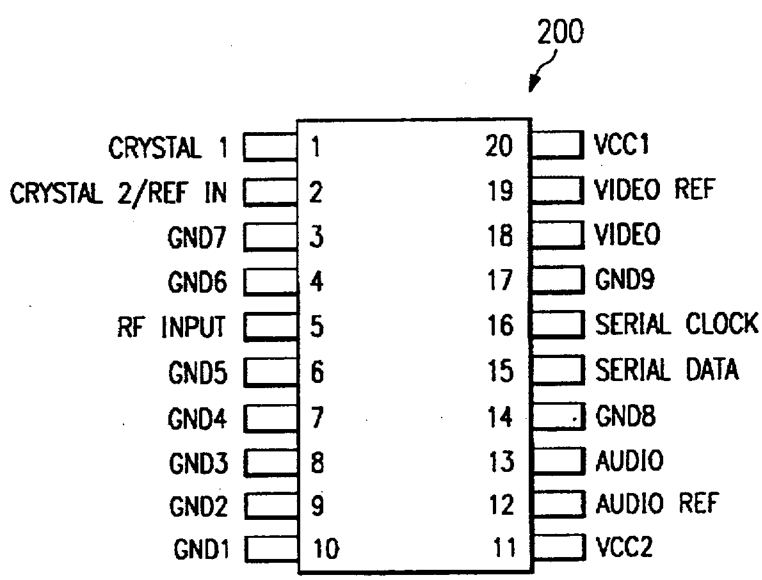
128 Claims, 4 Drawing Sheets





100
AGC
120
IF PROCESSOR
VSB VIDEO DETECTOR
VIDEO PROCESSOR AND DRIVER
VIDEO OUT
VIDEO REF
AFC
130
150
ANTENNA
RF CONVERTER
FM AUDIO DETECTOR
AUDIO PROCESSOR AND DRIVER
AUDIO OUT
AUDIO REF
110
170
140
180
160
190
SERIAL DATA
SERIAL CLOCK
SERIAL DIGITAL INTERFACE
BIAS AND CONTROL
FREQUENCY REFERENCE
CRYSTAL/REF

FIG. 1



200
CRYSTAL 1
CRYSTAL 2/REF IN
GND7
GND6
RF INPUT
GND5
GND4
GND3
GND2
GND1
1
2
3
4
5
6
7
8
9
10
20
19
18
17
16
15
14
13
12
11
VCC1
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VIDEO
GND9
SERIAL CLOCK
SERIAL DATA
GND8
AUDIO
AUDIO REF
VCC2

FIG. 2

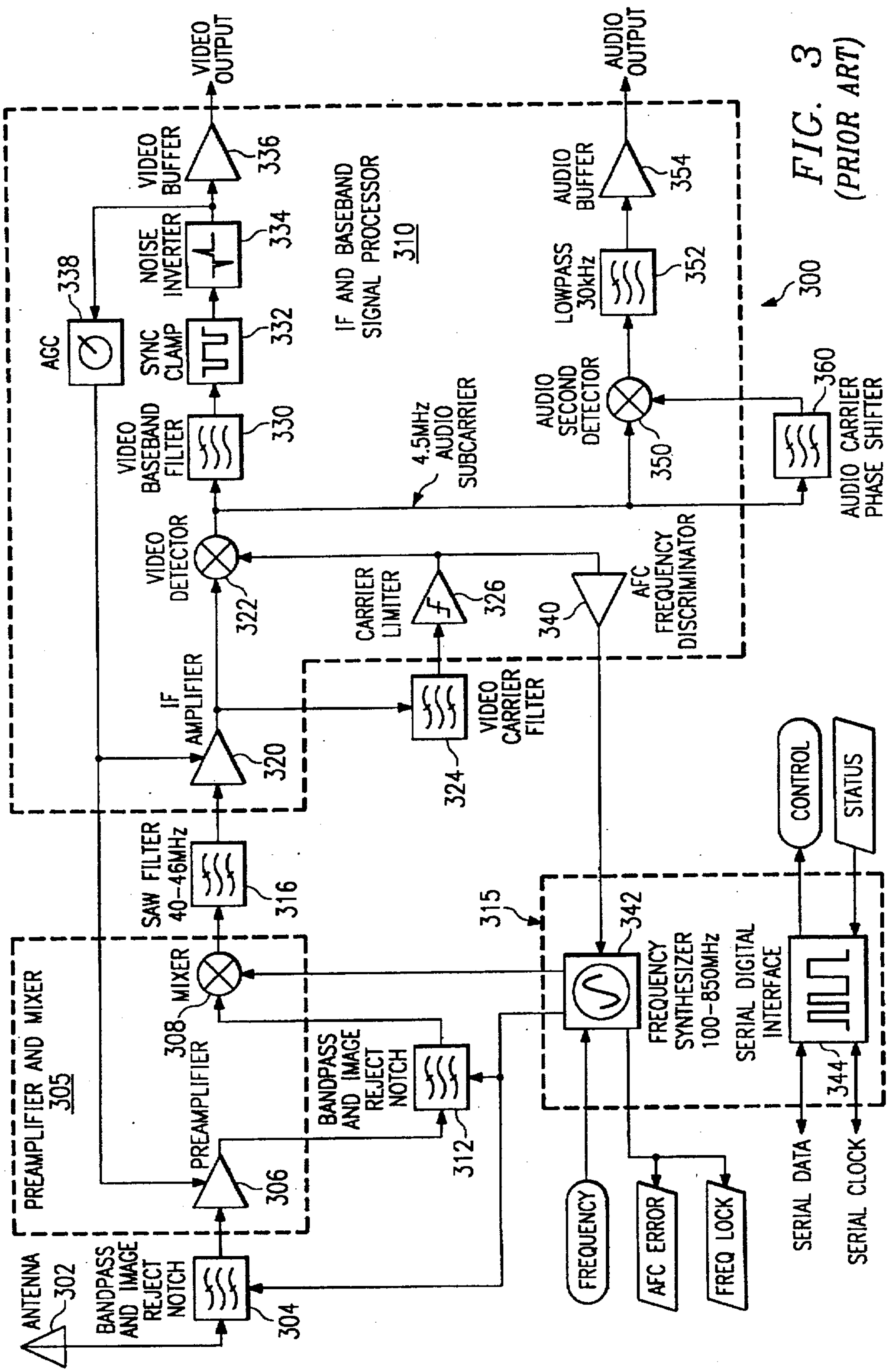


FIG. 3 (PRIOR ART)

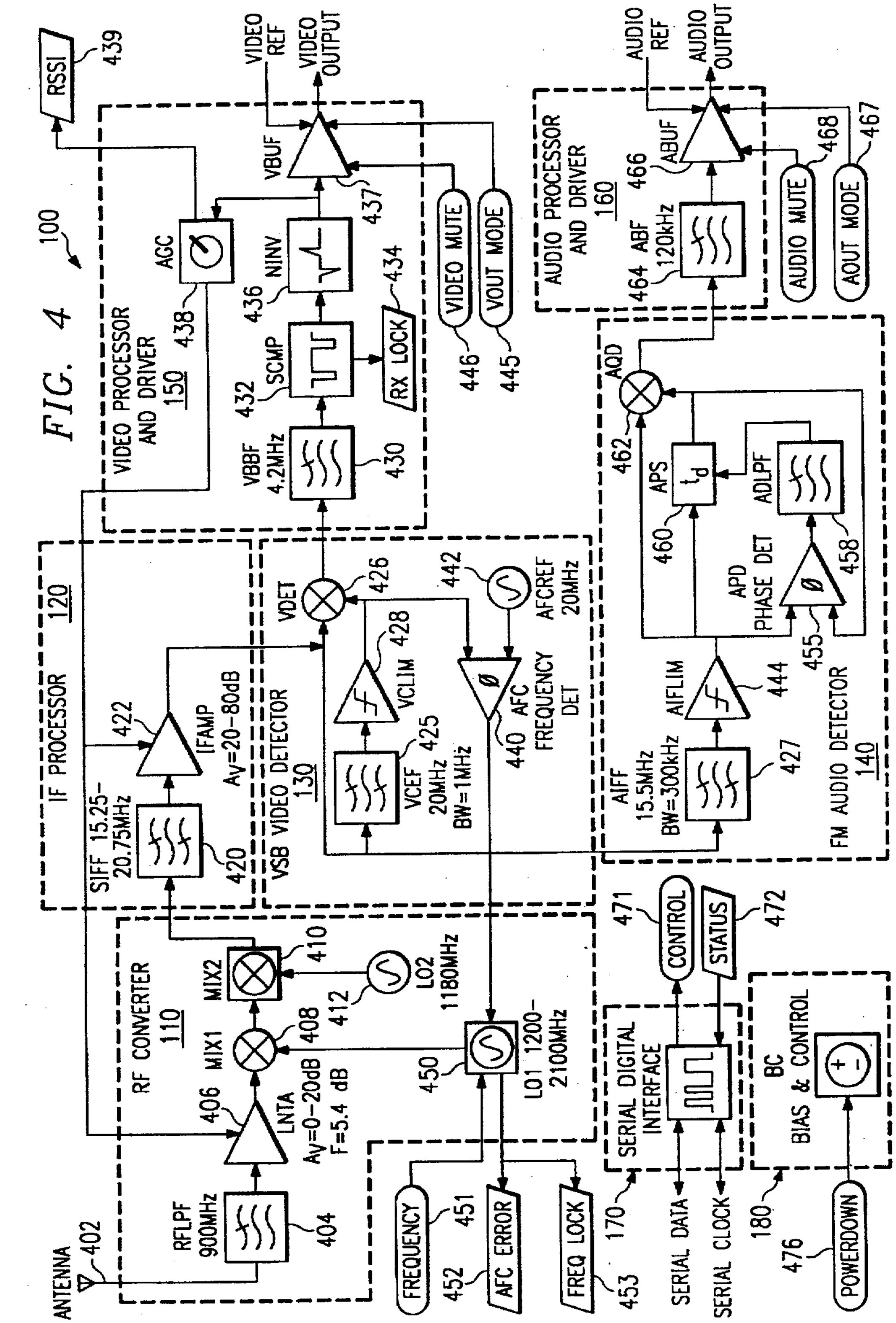
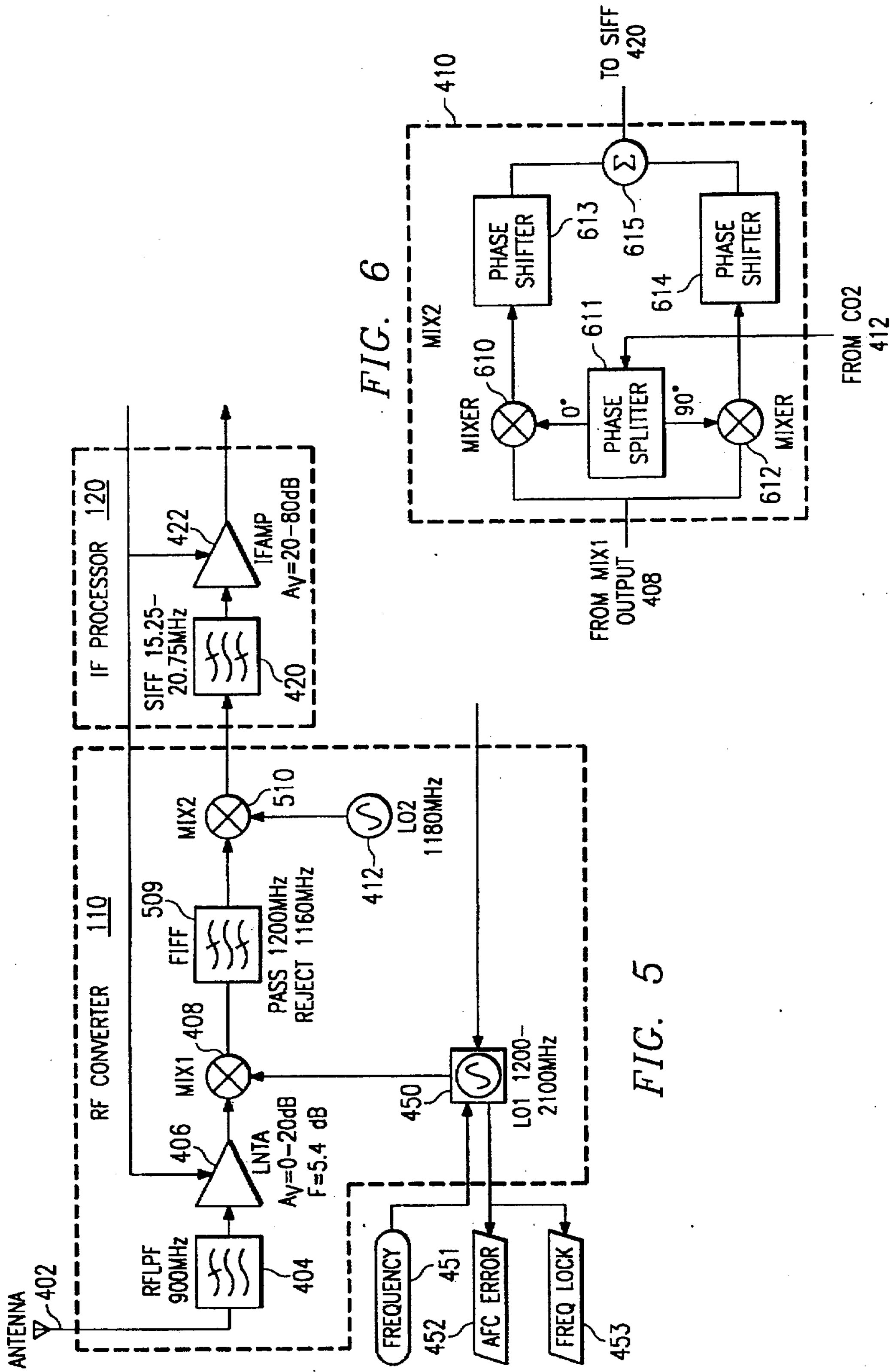

100
RSSI
439
VIDEO REF
VIDEO OUTPUT
VIDEO PROCESSOR AND DRIVER 150
AGC
438
VBUF
437
NINV
436
SCMP
432
RX LOCK
434
VBBF 4.2MHz
430
VIDEO MUTE
446
VOUT MODE
445
AUDIO REF
AUDIO OUTPUT
AUDIO PROCESSOR AND DRIVER 160
466
ABUF
464 ABF 120kHz
AUDIO MUTE
468
AOUT MODE
467
AQD
462
APS
t_d
460
ADLPF
APD PHASE DET
458
455
AIFLIM
444
AIFF 15.5MHz BW=300kHz
427
FM AUDIO DETECTOR 140
IF PROCESSOR 120
SIFF 15.25-20.75MHz
420
IFAMP Av=20-80dB
422
VSB VIDEO DETECTOR 130
VDET
426
VCLIM
428
VCEF 20MHz BW=1MHz
425
AFC FREQUENCY DET
440
AFCREF 20MHz
442
RF CONVERTER 110
MIX1
MIX2
410
408
LO2 1180MHz
412
LO1 1200-2100MHz
450
LNTA Av=0-20dB F=5.4 dB
406
RFLPF 900MHz
404
ANTENNA
402
FREQUENCY
451
AFC ERROR
452
FREQ LOCK
453
SERIAL DIGITAL INTERFACE
170
CONTROL
471
STATUS
472
SERIAL DATA
SERIAL CLOCK
BC BIAS & CONTROL
180
POWERDOWN
476

FIG. 4



410
TO SIFF 420
MIX2
PHASE SHIFTER
613
615
614
PHASE SHIFTER
611
MIXER
610
0°
PHASE SPLITTER
90°
MIXER
612
FROM CO2 412
FROM MIX1 OUTPUT 408
IF PROCESSOR 120
422
IFAMP Av=20-80dB
SIFF 15.25-20.75MHz
420
RF CONVERTER 110
509
FIFF
MIX2
510
LO2 1180MHz
412
PASS 1200MHz REJECT 1160MHz
408
MIX1
406
LNTA Av=0-20dB F=5.4 dB
450
LO1 1200-2100MHz
ANTENNA
402
RFLPF 900MHz
404
FREQUENCY
451
452
AFC ERROR
FREQ LOCK
453

FIG. 6

FIG. 5

HIGHLY INTEGRATED TELEVISION TUNER ON A SINGLE MICROCIRCUIT

TECHNICAL FIELD OF THE INVENTION

This invention relates to television tuner circuits and more particularly to a highly integrated television tuner fabricated in a single microcircuit device.

BACKGROUND OF THE INVENTION

One of the most significant costs in television manufacturing is the cost of the tuner. The typical cost of a television (TV) tuner is in the neighborhood of $15.00, which, relative to the cost of the entire television set, is very substantial. Part of the solution to reducing tuner cost is to reduce the number of components in the tuner.

Traditionally, tuners have been comprised of two basic components. The first component performs high frequency to intermediate frequency (RF to IF) conversion. Subsequently, the second component performs IF to baseband conversion. The TV tuner was originally designed for broadcast television reception within a television set, which is essentially a stand-alone unit containing a cathode ray picture tube. So, TV tuners were originally integral parts embedded in a single-purpose device.

Presently, however, state-of-the-art consumer electronic devices use TV tuners that are not a built-in part of a television set. The tuner is a separate element that is connected to a cathode ray picture tube at some point, but the tuner is not an integral part of the monitor. For example, TV tuners may be fabricated on circuit boards and then installed in personal computer (PC) systems, thereby allowing the PC to function as a television set. These tuners convert a radio frequency television signal into a baseband (or low frequency) video signal which can then be passed on to other elements in the PC for video processing applications.

The circuit component that performs the RF-to-IF conversion typically comprises one or two integrated circuits and numerous discrete elements—inductors, capacitors and/or transistors. The IF-to-baseband conversion typically includes another integrated circuit, several filter elements, such as ceramic filters and SAW filters, a series of tuning and control elements, such as resistors and potentiometers, variable inductors and/or capacitors, and some other additional external components. Thus, the complexity of the tuner is fairly high and typically there may be between 100 and 200 elements on a circuit board. Furthermore, state-of-the-art TV tuners still require that each tuner be aligned by manual tuning before leaving the factory. This manual tuning is one of the most expensive costs associated with the manufacturing process and an important factor in the cost of tuners.

Broadcast television tuners of the past have gone through an evolution over a period of more than 60 years. The earliest tuners utilized vacuum tube technology and required that the minimum number of vacuum tubes possible be used due to their cost, power consumption and dimensions. Therefore, passive components, such as resistors, capacitors, inductors and transformers, were used as much as possible in most designs. This style of design continued until about 1960 when TV tuner components, particularly vacuum tubess began to be replaced by bipolar and MOS transistors. However; the active device count still defined the cost and size limits of TV tuners and active device count minimization continued.

In the early 1970's the integrated circuit became viable as an element in the television tuner and the design techniques were dramatically changed. Many functions of the tuner utilizing only one tube or transistor were being replaced with 4 to 20 individual transistors which could perform the same function with better precision, less spaced less powers less heat generation and lower costa The introduction of the integrated circuit was gradual, first encompassing only low frequency elements and then eventually high frequency active elements. Nonetheless, many passive elements external to the integrated circuits remained in TV tuner designs.

One advance, the SAW (surface acoustic wave) filters made a significant change in that several manually tuned inductors and capacitors could be removed from the tuners and receive-filtering performance could be improved within a much smaller space and at reduced cost. However, the SAW filter, which is fabricated on a ceramic substrate, cannot be integrated on a silicon wafer with the rest of the active circuitry and must therefore remain a discrete component in the final design. The trend of the 1980's was to miniaturize all of the passive components and simplify their associated manual tuning at the factory. In recent years. TV tuners have been reduced in size from requiring fairly large enclosures, about 2"×5"×1", to much smaller enclosures, about ½"×2"×⅜". There is a high premium placed on small size because TV tuners are being used in smaller and smaller computers, television sets and VCRs. As the equipment in which tuners are used becomes smaller, the size of the TV tuner must decrease also.

As the size of the tuner goes down, and as tuners are used in a wider variety of devices, cost becomes more critical and must be reduced as much as possible in order not to represent a large portion of the final product cost. When a tuner is used in a television set, the tuner size is less critical because the television set inherently has a large mass. But when a tuner is used in other electronic equipments space becomes a premium and the footprint of the tuner becomes critical.

Accordingly, it is one object of the invention to provide a TV tuner which has a relatively low cost and a small footprint for use on a printed circuit board.

It is another object of the present invention to provide a TV tuner that meets or exceeds the performance of state-of-the-art TV tuners while at the same time reducing the number of external components needed, thereby decreasing the complexity of the printed circuit board and the amount of circuit board area needed by the TV tuner.

It is the further object of the present invention no allow for computer control of the TV tuner by a serial bus so that the TV tuner may be controlled by a microcontroller imbedded in the television set, personal computer, or other video device.

It is the further object of the present invention to provide a TV tuner with computer-controlled output impedance characteristics to accommodate different load specifications.

SUMMARY OF THE INVENTION

These and other problems have been solved by integrating a TV tuner that takes a broad band of frequencies as an input and performs video demodulation over a very narrow band width. To accomplish this, an architecture was chosen to perform an up-conversion of the RF input signal to a higher internal frequency, which allows the present invention to have minimal filtering on the input stages of the receiver. The present invention is therefore able to operate without variable tuned filtering on its input, with the exception of a single fixed tuned filter. This eliminates the need for precisely controlled variable tuned filters which must be

mechanically aligned during manufacture and are subject to variation in performance due to age, temperature, humidity, vibration and power supply performance. This was a critical drawback of previous tuners that had to be eliminated because it is a source of tremendous error and distortion, as well as complexity.

Furthermore, it is advantageous to have the up-conversion performed on-chip, because eliminating the inputting or outputting of high frequency signals to or from the integrated circuit of the present invention will avoid drive capability problems associated with high frequency signals and noise coupling problems associated with integrated circuit external interconnections.

The present invention next performs a down-conversion from the higher frequency using an image rejection mixing scheme that provides a tightly controlled down-conversion with very little added distortion. This also further minimizes the on-chip filtering effects by minimizing the amount of filtering that must be done at high frequency. The result is that high frequency operations within the present invention are limited to the initial stage inside the chip and are performed at a very low impedance level in order to minimize distortion. After the down conversions all remaining filtering is performed in continuous time mode filters that are realizable in an integrated circuit.

Another technical advantage of the present invention is that no manual-tuning of the final design is required. There are no external components needed to perform adjustments, as there is in the prior art.

A further technical advantage of the present invention is that the output impedance is selectable, based on system requirements, thereby minimizing power dissipation for certain applications. The present invention can drive a coaxial cable with a low output impedance driver or, for board level applications, the present invention can be configured to drive a high output impedance. The present invention can also match a very precise load impedance defined by a single external element.

The present invention advantageously utilizes much less board space than previous designs (on the order of 5% to 10% of the prior art designs) and dissipates only about one-fifth of the power. The present invention also advantageously operates on a single voltage level, as opposed to two or three levels for previous designs.

A further technical advantage of the present invention is that the need for a metal enclosure is reduced integration, by itself, allows for sufficient shielding to meet interference standards.

The monolithic television (MTV) tuner embodied in the present invention is intended to replace the TV tuner modules presently used in most broadcast television receiver devices. The MTV tuner integrates all of the functional elements of a tuner, except for the quartz crystal frequency reference and power supply bypass capacitors. The level of integration of the present invention dramatically reduces the cost of the basic TV tuner and enhances its manufacturability and reliability.

The TV tuner of the present invention is controlled externally by a computer or controller via a digital serial bus interface (I^2C). A preferred embodiment of the present invention provides an antenna input capable of being connected directly to a standard coaxial cable, thereby allowing both antenna and cable television applications. Furthermore, the baseband video and audio outputs are designed for high impedance or low impedance applications. The high impedance mode is a driver for short interconnections on a printed circuit and gives the benefit of low operating power. The low impedance mode drives an industry standard studio cable interface requiring greater supply power.

A preferred embodiment of the present invention is designed to operate on frequencies used for both over-the-air broadcasts and cable television with National Television Standards Committee (NTSC) encoded video. Receiver sensitivity is set to be limited by the antenna noise temperature for VHF systems. The present invention also employs a wide-range automatic gain control (AGC).

The baseband video output of the present invention is leveled, or has minimal variation in video amplitude with respect to antenna RF signal level, and can be configured via software for either low impedance studio standard loads or for high impedance loads with lower operating current. Audio output is broadband composite to allow connection to an external MTS decoder. The audio output can be configured for low impedance studio standard loads or high impedance loads via software control.

Control is accomplished via an I^2C bus interface. The bias and control circuits in a preferred embodiment of the present invention contain internal registers which can be updated via the control bus for operating frequency, video and audio transmission standards such as NTSC, PAL, SECAM and MTS, power, and test modes. Status of the bias and control circuits can be checked via a status register accessible through the I^2C bus interface. Status data include AFC error, channel lock and received signal strength indicator.

The operating frequency of the present invention is referenced to an external crystal or reference frequency generator. A minimum of external components are used with the present invention and no tuning of any components is required.

The present invention may be implemented in Bipolar, BiCMOS, or CMOS processes, however, a preferred embodiment of the present invention employs a BiCMOS process to reduce the difficulty in developing the design by allowing maximum flexibility.

A preferred embodiment of the present invention performs sound processing after IF amplification and prior to video demodulation. This method is known as "quasi-split" sound and is not used in the majority of television receivers for cost reasons. Quasi-split sound does not have a "buzz" in the audio signal during periods when the video picture is very white. In the integrated circuit of the present invention, quasi-split sound is a negligibly small cost and thus may be integrated in all embodiments of the present invention, integration of quasi-split sound further reduces component count and does not require external manual adjustments.

The foregoing has outlined rather broadly the features and technical advantages of the present invention in order that the detailed description of the integrated television tuner that follows may be better understood. Additional features and advantages of the monolithic television tuner will be described hereinafter which form the subject of the claims of the invention. It should be appreciated by those skilled in the art that the conception and the specific embodiment disclosed may be readily utilized as a basis for modifying or designing other structures for carrying out the same purposes of the present invention. It should also be realized by those skilled in the art that such equivalent constructions do not depart from the spirit and scope of the invention as set forth in the appended claims.

BRIEF DESCRIPTION OF THE DRAWINGS

For a more complete understanding of the present invention, and the advantages thereof, reference is now

made to the following descriptions taken in conjunction with the accompanying drawings, in which:

FIG. 1 is a high level block diagram of the functional components of the present invention;

FIG. 2 is a pin layout diagram of an integrated circuit in accordance with the present invention;

FIG. 3 is a detailed block diagram of a state-of-the-art television tuner found in the prior art;

FIG. 4 is a detailed block diagram of the circuit components of the present invention;

FIG. 5 shows an alternate embodiment of the RF front-end of the present invention; and

FIG. 6 is a detailed block diagram of MIX2 of FIG. 4.

DETAILED DESCRIPTION OF THE PRIOR ART

Before discussing the monolithic television tuner of the present invention, it will be useful to discuss a state-of-the-art television tuner found in the prior art.

While there have been theoretical proposals to integrate TV tuners in a single microcircuit, none are known to have been implemented. The next best definition of the known prior art, then, is a highly miniaturized, but not fully integrated, tuner as shown in FIG. 3.

FIG. 3 depicts a functional electrical block diagram of a present state-of-the-art TV tuner configuration. Television tuner **300** is constructed in a single metallically shielded assembly containing a printed circuit board on which all of the associated tuner components are mounted and electrically connected. TV tuner **300** is designed to be a module mounted on other printed circuit boards to allow for direct connection of the input and output signals to their appropriate terminations within the television receiving system. The metal shield is used to keep undesired external signals from interfering with the operation of the TV tuner **300** and to prevent TV tuner **300** from radiating signals that interfere with the operation of external devices.

Prior art television tuner **300** is comprised of three integrated circuits: preamplifier and mixer **305**, IF and baseband signal processor **310** and frequency synthesizer and Inter integrated Circuit (IIC or I^{2}C) bus interface **315**. Television tuner **300** is also comprised of a plurality of discrete components, including bandpass and image reject notch filter **304**, bandpass and image reject notch filter **312**, surface acoustic wave (SAW) filter **316**, video carrier filter **324**, and audio carrier phase shifter **360**.

Television tuner **300** receives a standard television RF signal from either antenna **302** or a cable system connection (not shown) through bandpass and image reject notch filter **304**. Bandpass and image reject notch filter **304** limits the signals entering TV tuner **300** so that a minimum number of undesired signals exist in TV tuner **300**. Filter **304** therefore limits the image response caused by the first mixer, described later. Filter **304** also attenuates signals not in a fairly narrow (100 MHz) range about the desired signal. Finally, known interference signals, such as FM broadcast, shortwave service signals, signals in the intermediate frequency band and Citizen Band radio signals, are specifically rejected by filter **304**.

Preamplifier **306** of preamplifier and mixer **305** receives the output of bandpass and image reject notch filter **304** and raises the signal level (10 dB) with minimum increase in the noise level (typically 8–10 dB). The gain of preamplifier **306** is controlled by automatic gain control (AGC) **338**, so that when a very strong signal enters TV tuner **300**, overall gain is reduced, resulting in less distortion in the preamplifier than without the gain reduction.

The output of preamplifier **306** is sent to bandpass and image reject notch filter **312**, with the same basic requirement of minimizing the passage of potential interference signals. Filter **312** is external to preamplifier and mixer **305** and is comprised of a plurality of discrete elements, including capacitors, inductors and varactor diodes.

The output of bandpass and image reject notch filter **312** is then sent back to mixer **308** in preamplifier and mixer **305**. Mixer **308** mixes the output of filter **312** with the output of a local oscillator, frequency synthesizer **342**, which has a frequency chosen to be higher than the desired receiver carrier by 45.75 MHz, Thus, the output of mixer **308** is 45.75 MHz. There also is an image signal due to mixer **308** at 91.5 MHz above the input frequency, which is removed by filter **304** and filter **312**. Therefore, as the frequency of frequency synthesizer **342** is tuned to receive signals of different carrier frequencies, the bandpass and image reject filters **304** and **312** must also be tuned to properly pass only the desired signals and not the mixer images.

Frequency synthesizer **342** receives an input frequency reference signal (usually 16 bits) and outputs the status signals AUTOMATIC FREQUENCY CONTROL (AFC) ERROR and FREQUENCY (FREQ) LOCK. Additionally, a tuning signal which is used by the voltage controlled oscillator (VCO) in frequency synthesizer **342** is output from frequency synthesizer **342** to bandpass and image reject notch filters **304** and **312**. A local oscillator signal is output from frequency synthesizer **342** to mixer **308**.

The 45.75 MHz output signal of mixer **308** then passes through SAW (surface acoustic wave) filter **316**, which limits the bandwidth of the signal to only one (1) channel (6 MHz for NTSC standard) and applies a linear attenuation in frequency known as the Nyquist slope around the visual carrier frequency. The linear attenuation by SAW filter **316** converts the signal from a vestigial sideband signal to one which is equivalent to a single sideband with a carriers, so that the frequency response of the signal after demodulation is flat over the video bandwidth. SAW filter **316** is very "lossy" (on the order of 25 dB), so the input to SAW filter **316** is amplified by a preamplifier (not shown) by a corresponding amount to minimize noise effects.

The output of SAW filter **316** is input to intermediate frequency (IF) amplifier **320** in IF and baseband signal processor **310**. IF amplifier **320** provides most of the overall gain of TV tuner **300** and receives gain control from AGC **338**.

The output of IF amplifier **320** is sent to video detector **322** and is also sent off-chip to external video carrier filter **324**. This is the stage at which video demodulation is performed. Video detector **322** is essentially a mixer with the local oscillator input connected to the output of video carrier filter **324** through carrier amplitude limiter **326**. The output of the carrier limiter **326** is an in-phase representation of the video carrier signal without any modulation applied to it. The output of carrier limiter **326** is received by video detector **322**, which mixes the output of carrier limiter **326** with the output of IF amplifier **320**.

AFC frequency discriminator **340** is used in the prior art device to detect the difference between the carrier frequency contained in the output of carrier limiter **326** and a known valid carrier frequency reference. The output signal on the output of AFC frequency discriminator **340** is an error signal which is used to drive frequency synthesizer **342** in a direction that will reduce the error between the output of carrier limiter **326** and the known valid carrier frequency reference.

The output of the video detector **322** is a baseband video signal combined with several high frequency mixing artifacts. These artifacts are removed by a video baseband filter **330**. The output of video baseband filter **330** is fed to synchronization pulse clamp (sync clamp) **332**, which sets the level of the sync pulses to a standard level.

Next, the output of sync clamp **332** is sent to noise inverter **334**, which removes large noise spikes from the signal. The output of noise inverter **334** is sent to video buffer **336**, which is configured to drive fairly high circuit board impedances of approximately 1000 to 2000 ohms.

The output of noise inverter **334** is also sent to AGC (automatic gain control) **338**, which compares the level of the synchronization pulses to the signal blanking level to measure the incoming signal strength and generates a gain control signal which is used by IF amplifier **320** and RF preamplifier **306** to dynamically adjust the gain of the TV tuner **300** for the correct level at the final output.

The audio signal is an FM signal which follows the same path as the video through video detector **322**. At the output of video detector **322**, the audio signal appears as a subcarrier at 4.5 MHz, due to the fact that the audio signal comes into prior art TV tuner **300** 4.5 MHz higher in frequency than the desired video carrier. The audio subcarrier is passed on to an FM quadrature demodulator. The FM quadrature demodulator is comprised of a mixer, audio second detector **350**, and a 90 degree (at 4.5 MHz) phase shifter, audio carrier phase shifter **360**. The output of the audio second detector **350** is a baseband audio signal, which is filtered by lowpass (30 kHz) filter **352** to remove any undesired high frequency components. The output of lowpass filter **352** is finally passed on to audio buffer **354**, which drives an audio amplifier that ultimately drives a speaker. Serial digital interface **344** receives SERIAL DATA and SERIAL CLOCK inputs to provide control and update status for the prior art television receiver.

Baseband and image reject notch filters **304** and **312** are typically comprised of a plurality of capacitors, inductors and varactor diodes. Video carrier filter **324** is usually comprised of three discrete elements: an inductor and two capacitors. Likewise, audio carrier phase shifter **360** is also comprised of an inductor and two capacitors. In addition to the circuit elements shown as discrete components outside of circuit elements **305**, **310** and **315** in FIG. **3**, other discrete components (not shown) are connected to IF and baseband signal processor **310** and frequency synthesizer **342** for tuning purposes. Frequency synthesizer **342** is typically tuned by several external capacitors, inductors and/or varactor diodes. Video buffer **336** and audio buffer **354** will also typically employ external discrete elements, such as resistors, capacitors and/or transistors. Video baseband filter **330** and low pass filter **352** may also employ external inductors and capacitors.

DETAILED DESCRIPTION OF THE INVENTION

FIG. **1** depicts a high-level block diagram of the functional elements contained in a preferred embodiment of TV tuner **100**. The RF signal received from the antenna or other source is converted by RF converter **110** to an intermediate frequency (IF) signal and sent to IF processor **120**. The output of IF processor **120** is sent to Vestigial Sideband (VSB) detector **130** and video processor and driver **150** to produce the video output signal of the present invention. The video signal is detected via a synchronous AM demodulator.

The output of IF processor **120** is also sent to FM detector **140** and audio processor and driver **160** in order to produce the audio output signal of the present invention. Audio detection is performed by a quadrature FM detector utilizing a phase-locked loop for the quadrature reference. I^2C interface **170** receives I^2C DATA and I^2C CLOCK signals from the I^2C interface bus. Frequency reference **190** is an oscillator that is synchronized to the external reference crystal. Bias and control logic circuit **180** defines internal bias voltages and currents and maintains the status and control registers of the television tuner of the present invention.

FIG. **2** illustrates a pin layout diagram of a preferred embodiment of an integrated circuit in accordance with the present invention. Integrated circuit **200** is connected to a supply voltage on pins **11** and **20** (VCC1 and VCC2) and is grounded on pins **3**, **4**, **6–10**, **14** and **17** (GND1–GND9). An input reference signal from an external crystal oscillator is supplied on pins **1** and **2** (CRYSTAL **1** and CRYSTAL 2/REF IN). Integrated circuit **200** is connected to an IIC (I^2C) interface bus on pin **16** (SERIAL CLOCK) and pin **15** (SERIAL DATA). The RF input from an antenna or other source is connected to pin **5** (RF INPUT). Integrated circuit **200** outputs video and audio signals on pins **18** and **19** (VIDEO and VIDEO REF) and pins **12** and **13** (AUDIO and AUDIO REF).

FIG. **4** is a detailed electrical block diagram of TV tuner **100** in accordance with a preferred embodiment of the present invention. FIG. **4** depicts an up-converting dual conversion superheterodyne receiver. The Vestigial Sideband (VSB) encoded video signal is processed via a Nyquist slope receiver attenuation characteristic filter prior to detection.

The RF signal enters TV tuner **100** from the antenna **402** (or cable, not shown) and is passed through RF low-pass filter (RFLPF) **404** to limit the incoming band to below 900 MHz. The filtered RF signal is amplified up to 20 dB by a gain-controlled low-noise transconductance amplifier (LNTA) **406**.

The input signal received by antenna **402**, filtered by RFLPF **404** and amplified by LNTA **406** is the standard broadcast television spectrum. Channels **2–13** in the VHF band extend from 54 Megahertz to 216 MHz and channels **14–83** in the UHF band extend from 410 MHz to 890 MHz. Each channel in the VHF and UHF band has a 6 MHZ bandwidth with the video carrier frequency located 1.25 MHz above the lower band edges a color carrier frequency located 3.58 MHz above the video carrier, and an audio carrier frequency located 4.5 MHz above the video carrier. For example, Channel 2 has a 6 MHz bandwidth from 54 to 60 MHz, a video carrier at 55.25 MHz, a color subcarrier at 58.83 MHz and an audio carrier at 59.75 MHz.

The output of first local oscillator (LO1) **450**, operating between 1200 and 2100 MHz, is mixed in first mixer (MIX1) **408** with the RF signal to generate a first IF video carrier frequency of 1200 MHz. This approach leads to minimum distortion due to mixer images and harmonic mixing. The first IF is crudely filtered by the bandwidth limitation of first mixer **408** to minimize harmonic effects.

In a basic mixer scheme, a mixer receives two inputs, an RF input and a local oscillator (LO) input, and generates an IF output. The RF input has the general form:

$$V_{RF}=A\cos(\omega_{RF}t), \qquad [1]$$

the local oscillator input has the general form:

$$V_{LO}=B\cos(\omega_{LO}t), \qquad [2]$$

and the IF output is given by:

$$V_{IF} = V_{RF} \cdot V_{LO} = AB/2[\cos(\omega_{RF}-\omega_{LO})t + \cos(\omega_{RF}+\omega_{LO})t]. \quad [3]$$

In the frequency domain, the IF frequencies are given by: $f_{IF}=|f_{RF}\pm f_{LO}|$. In realizable mixers, there are various non-linearities which cause the generation of harmonics at multiples of the RF and IF frequencies. Thus, the IF spectrum contains harmonics at $f_{IF}=mf_{RF}\pm nf_{LO}$, where m and n are integers.

For example, if the output of local oscillator **450** is 400 MHz and the desired input signal to be demodulated from the antenna is 800 MHz, a basic mixer would perform an additive mix to obtain a 1200 MHz output at the output of mixer **408**. However, since frequency synthesizer **450** also has a harmonic at 800 MHz, and a signal exists on the antenna at 400 MHz, these signals are also added to produce an 800 MHz output that would interfere with the desired signal to be demodulated at 800 MHz.

Mixer **408** is a subtractive mixer that will subtract the frequency of local oscillator **450** from the input RF spectrum between 0 Hz to 900 MHz that is filtered by RFLPF **404**. For example, if the desired frequency to be demodulated in the RF input is 400 MHz, the local oscillator would be set to 1600 MHz and a basic mixer would generate two IF outputs:

$$f_{IF}=|400-1600|=1200 \text{ MHz and} \quad [4]$$

$$f_{IF}=|400+1600|=2000 \text{ MHz} \quad [5]$$

Since mixer **408** is a subtractive mixers the 2000 MHz signal will be filtered out and only the 1200 MHz signal will be passed on to mixer **410**. Also, since mixer **408** is a subtractive mixer, the incoming spectrum is inverted on the output, so than the video carrier is 1.25 MHz below the upper edge of the 6 MHz channel bandwidth, the color carrier is 3.58 MHz below the video carrier and the aural carrier is 4.5 MHz below the video carrier. For example, the 6 MHz bandwidth of channel **2** from 54–60 MHz would appear inverted between 1201.25 MHz and 1195.25 MHz. The channel **2** video carrier would appear at 1200 MHz, the color carrier would appear at 1196.42 and the aural carrier would appear at 1195.5 MHz.

As noted above, it is ms a characteristic of both the input RF signal received, by mixer **408** from LNTA **406** and of frequency synthesizer **405** that harmonic signals exist at multiples of the desired frequencies. By using a subtractive mixer and a sufficiently high-frequency output from oscillator **450**, the harmonics on the output of frequency synthesizer **450** will all be more than 1200 MHz higher in frequency than any of the input frequencies existing in the 0 Hz to 900 MHz range from antenna **402**.

For example, the lowest output of local oscillator 450 is 1200 MHz. The first harmonic of this signal exists an 2400 MHz. Since the highest output signal passing through LNTA **406** is 900 MHz, the subtractive mix of the 2400 MHz harmonic and the 900 MHz signal from antenna **402** would yield a signal of |900–1200|=1300 MHz. This is higher than the desired 1200 MHz output from mixer **408**, and may therefore be filtered out by low-pass filtering. Using a local oscillator whose output is higher that the RF input frequency is known as high-side injection.

The first IF signal of 1200 MHz is mixed in second mixer (MIX2) **410**, which is an image-rejection mixer, with the fixed 1180 MHz reference output of second local oscillator (LO2) **412** to generate the second IF at 20 MHz visual carrier. Because the RF input signal is lower in frequency than the LO referenced, the mixing of the two signals will result in a down conversion of the RF input. The IF frequencies, given by $f_{IF}=|f_{RF}\pm f_{LO}|$, have carriers at f_{IF}=1200–1180=20 MHZ and at f_{IF}=1200+1180=2380 MHZ. The image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF}=|1160–1180|=20 MHz.

The image rejection mixer receives an input signal from MIX1 that has the general form:

$$V_{RF}=A\cos(\omega_{RF}t), \quad [6]$$

and the local oscillator signal from LO2 is split into two phase quadrature signals having the general form:

$$V_{LOI}=B\cos(\omega_{LO}t) \text{ and} \quad [7]$$

$$V_{LOQ}=B\sin(\omega_{LO}t). \quad [8]$$

The desired output signal, V_{IF}, resulting from the mixing of the RF input and the phase quadrature signals has the general form:

$$V_{IF}=AB\cos(\omega_{RF}-\omega_{LO})t. \quad [9]$$

The desired mixer signal at $\omega_{IF}=\omega_{RF}-\omega_{LO}$ appears in the mixer outputs while the undesired mixer image signal at $\omega_{IF}=\omega_{RF}+\omega_{LO}$ (−1160+1180=20) is rejected.

FIG. 5 depicts an alternate embodiment of the RF front-end (i.e., RF converter **110** and IF processor **120**) of the present invention. First IF filter (FIFF) **509** between mixer **408** and mixer **510** filters out all frequency components of the image falling on 1160 MHz, such that mixer **510** is a basic mixer similar to mixer **408**, rather than an image rejection mixer. While the image rejection mixer is easier to integrate, it has accuracy limitations, and noise and distortion figures may suffer.

Returning to FIG. **4**, the second IF signal is next filtered for the final video bandwidth by second IF filter (SIFF) **420**, a low-distortion continuous-time bandpass filter. SIFF **420** also comprises an automatically tuned low-pass Nyquist slope filter (NSF) for VSB balancing. The −6 dB point on the slope of the NSF is maintained at the visual carrier frequency (20 MHz).

The output of SIFF **420** is coupled to IF amplifier (IFAMP) **422**, a gain-controlled amplifier which applies up to 80 dB of gain. The output of IFAMP **422** is then mixed in video detector (VDET) **426** with the video carrier, which has been extracted by video extraction filter (VCEF) **425** and video carrier limiter (VCLIM) **428** to produce the baseband video output.

The video carrier is compared by automatic frequency control (AFC, frequency detector (DET) **440** with a synthesized 20 MHz reference signal from AFCREF **442** to generate a frequency error signal which is passed back to LO1 for automatic frequency control.

The baseband video from VDET **426** is passed through a low-pass video bandwidth filter (VBBF) **430** to remove detection harmonics, The synchronization pulses are clamped to a reference level by sync clamp (SCMP) **432**, which contains a phase-locked loop that is locked to the horizontal sync rate, RX LOCK **434**, or vertical rate as an option, to control the clamp timing. Timing is also generated in SCMP 432 for AGC 438 and NINV **436**. SCMP **432** amplifies the video to its output voltage level by a fixed gain of 20 dB. The clamped video is processed by noise inverter (NINV) **436** to remove excessively large positive or negative amplitude noise pulses.

The resulting video is tested for blanking to sync level by AGC **438**, which adjusts the gain of LNTA **406** and IFAMP

422 to maintain a standard 1 Volt peak-to-peak video signal. The video at this stage is buffered by video buffer (VBUF) 437 for off-chip drive using one of three signal impedance standards. VBUF 437 can also mute the signal.

Audio processing is accomplished by extracting the audio signal at the output of IFAMP 422 through narrow bandpass audio IF filter (AIFF) 427. AIFF 427 produces an FM signal that is limited by audio IF limiter (AIFLIM) 444 to remove AM noise and distortion.

A delay-locked loop comprised of audio phase detector (APD) 455, audio delay-locked loop low-pass filter (ADLPF) 458, and audio phase shifter (APS) 460 generates a signal that is delayed by 90° at the FM carrier frequency and mixes that signal with the limited FM signal in audio quadrature detector (AQD) 462 for quadrature detection of the FM. The audio baseband is passed through a 120 kHz low-pass audio baseband filter (ABF) 464 to remove detection artifacts. The audio is buffered in audio buffer (ABUF) 466 for off-chip drive with selectable drive impedances at a 0.35 Volt (rms) level.

Control registers 471 and status registers 472 are interfaced to serial digital interface 170. Serial digital interface 170 in turn receives external commands from an external microprocessor or microcontroller through SERIAL DATA and SERIAL CLOCK control lines. In a preferred embodiment of the present invention, serial digital interface 170 may be a Inter integrated Circuit (IIC or I²C) interface, which is a proprietary specification of Philips Corporation.

System Block Specifications

The overall design of TV tuner 100 is guided by the limitations of what can be reliably designed on an integrated circuit. Blocks of TV tuner 100 are specified by referring to actual circuit elements so that capabilities such as noise, maximum signal handling, intermodulation and gain can be ascertained. Perhaps, the most important element of such a design is to limit expected chip performance so that it can be manufactured with a high yield and, therefore, a low cost. Note that all measurements in dBm are referenced to a standard video RF and baseband impedance of 75 Ω.

RF Low-Pass Filter (RFLPF)

Table 1 lists the operating parameters of RF low-pass filter (RFLPF) 404 in accordance with a preferred embodiment of the present invention. RFLPF 404 is a 900 MHz third order low-pass filter comprising package lead inductance, printed circuit capacitance and on-chip capacitors. For most applications RFLPF 404 should be satisfactory for image (2400–3300 MHz) rejection in MIX1. External filtering matched to 75 ohm impedance could be added for extreme situations, such as very large out-of-band signals.

TABLE 1

RFLPF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	−1		0	dB
f_c	Cutoff Frequency (−3 dB)		900		MHz

Low Noise Transconductance Amplifier (LNTA)

Table 2 lists the operating parameters of low-noise transconductance amplifier (LNTA) 406 in accordance with a preferred embodiment of the present invention. LNTA 406 is the front end of the receiver. The single-ended input signal is converted to a differential current which will be passed through a mixer chain. The critical function of this circuit is to maintain a very low noise figure (NF), a substantial gain (G_V), and a high third-order intermodulation intercept (IP_3). The gain is determined relative to the rest of the circuit so that the noise figure of this block dominates the total receiver noise figure. Note that, since this is a transconductance amplifier, the gain is defined by the load impedance of the final mixer stage and must be properly defined before a final determination of LNTA 406 performance. AGC capability is included to reduce the impact of distortion, such as intermodulation or crossmodulation, in the presence of high level signals, The input is impedance matched to 75 Ω.

TABLE 2

LNTA Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain - no AGC	20		30	dB
G_{VAGC}	Voltage Gain - full AGC	−10		0	dB
NF	Noise Figure	4.5		5.4	dB
P_{-1dB}	1 dB Gain Compression Input Level	−24			dBm
IP_3	Third Order Input Intercept	−10			dBm
f_{in}	Input Frequency	40		900	MHz
I_{DC}	DC Bias Current		20		mA

First Mixer (MIX1)

Table 3 lists the operating parameters of first mixer (MIX1) 408 in accordance with a preferred embodiment of the present invention. Proper rejection of any undesired images and spurious signals due to local oscillator harmonics requires that MIX1 be configured for an up-conversion scheme with high-side injection. Thus, the output of MIX1 for the visual carrier is at the microwave frequency of 1200 MHz, the audio carrier is at 1195.5 MHz, and local oscillator 450 input runs between 1.2 and 2.1 GHz. A single-pole low-pass element is designed into the output circuitry of MIX1 to assist in the rejection of any down-converted signals due to local oscillator harmonics. The noise and distortion effects of MIX1 are included in LNTA 406 specifications. The design will include a bond-wire parallel LC network option to further band-limit any image energy at the output of this mixer if necessary.

TABLE 3

MIX1 Specifications

Parameter	Description	Min	Typ	Max	Unit
f_{RF}	RF Input Frequency	40		900	MHz
f_{LO}	LO1 Input Frequency	1200		2100	MHz
f_{IF}	IF Output Frequency	1195		1205	MHz
V_{LO}	LO1 Input Voltage		0.2		V(rms)

Second Mixer (MIX2)

Table 4 lists the operating parameters of second mixer (MIX2) 410 in accordance with a preferred embodiment of the present invention. The first IF of 1.2 GHz (visual carrier) is input to MIX2, which is an image-rejection mixer, which down-converts to a 20 MHz second IF. The image reaction is done to negate the need for filtering of the first mixer output in order to reject signals falling on the image at 1160 MHz from being down-converted by MIX2 if MIX2 were

just an ordinary subtractive mixer. An important feature of MIX2 is that it must achieve a high image rejection (at least 50 dB) without the need for post-fabrication trimming techniques to calibrate MIX2. As shown in FIG. 6, the 50 dB image rejection may be achieved by paralleling two Gilbert cell multipliers **610** and **612** at MIX2 input and using fairly high-order balanced RC-phase shifters **613**, **614** with very low component sensitivity with resulting phase errors via summer **615** of less than 0.2 degrees over a 20 MHz bandwidth. Multipliers **610** and **612** receive inputs from Phase Splitter **611** and MIX1 **408**. These are traditionally avoided due to relatively high noise levels, however, in a preferred embodiment of TV tuner **100**, the preious stages provide sufficient gain reduce the noise impact.

TABLE 4

MIX2 Specifications

Parameter	Description	Min	Typ	Max	Unit
G_I	Current Gain		0		dB
NF	Noise Figure			10	dB
f_{RF}	RF Input Frequency	1195		1205	MHz
f_{LO}	LO2 Input Frequency		1180		MHz
f_{IF}	IF Output Frequency	15.25		21.25	MHz
V_{LO}	LO2 Input Valtage		0.2		V(rms)
A_{IM}	Input Image Rejection	60			dB

Second IF Filter (SIFF)

Table 5 lists the operating parameters of second IF filter (SIFF) **420** in accordance with a preferred embodiment of the present invention. The 20 MHz IF signal is filtered for channel bandwidth in this integrated transconductor-capacitor (G_m-C) filter. A transitional Gaussian bandpass filter is used to provide adjacent channel and noise rejection. SIFF **420** keeps the distortions due to noise and out-of-band signals below −54 dBc. The in-channel group delay is to be maximally flat to the color subcarrier frequency. The aural signal passes through SIFF **420**. Since the up-conversion scheme inverts the channel frequencies at the output of MIX2, the visual carrier is now at 20 MHz and the aural carrier is at 15.5 MHz. The adjacent visual carrier is now at 14 MHz and the adjacent aural carrier is at 21.5 MHz leading to the filter specifications given below.

TABLE 5

SIFF Specifications

Parameter	Description	Min	Typ	Max	Unit
f_{p1}	Lower Passband Limit $A(\omega) < 0.5$ dB		16.75		MHz
f_{ph}	Upper Passband Limit $A(\omega) < 0.5$ dB		19.25		MHz
f_{s1}	Lower Stopband Limit $A(\omega) > 54$ dB		15.25		MHz
f_{sh}	Upper Stopband Limit $A(\omega) > 54$ dB		20.75		MHz
G_V	Voltage Gain		0		dB
IP_3	Third Order Input Intercept	20			dBm
NF	Noise Figure			15	dB

IF Amplifier (IFAMP)

Table 6 lists the operating parameters of IF amplifier (IFAMP) **422** in accordance with a preferred embodiment of the present invention. The 20 MHz signal is amplified by 80 dB in this block. The gain can be reduced 0 dB by application of an automatic gain control (AGC) signal voltage. The AGC is implemented in fully differential form for best on-chip noise rejection.

TABLE 6

IFAMP Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain - No AGC		80		dB
G_V	Voltage Gain - Full AGC		20		dB
NF	Noise Figure			10	dB
BW	Bandwidth	25			MHz

Visual Carrier Extraction Filter (VCEF)

Table 7 lists the operating parameters of visual carrier extraction filter (VCEF) **425** in accordance with a preferred embodiment of the present invention. The video carrier at 15 MHz is extracted from IFAMP **422** output for synchronous detection processing. VCEF **425** passes along any incidental carrier phase modulation (ICPM) in the Nyquist filter without any group delay distortion for best detector performance.

TABLE 7

VCEF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_0	Center Frequency		20		MHz
BW	Bandwidth	0.5	1	3	MHz
NF	Noise Figure			5	dB

Visual Carrier Limiter (VCLIM)

Table 8 lists the operating parameters of visual carrier limiter (VCLIM) **428** in accordance with a preferred embodiment of the present invention. The synchronous detector requires a fixed amplitude reference carrier. This is derived via VCEF **425** and then passed through VCLIM **428**. With input level variations of 20 dB and depth of modulation possibly exceeding 95 percent, VCLIM **428** output is stable within an accuracy of −50 dB. The circuit delay of VCEF **425** and VCLIM **428** combined must not exceed 45 degrees at the visual carrier frequency so that the detector output level and noise behavior is not degraded.

TABLE 8

VCLIM Specitications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	46			dB
V_{th}	Limiter Input Threshold		10		mV(p—p)
V_{limit}	Limiter Output Level	2			V(p—p)
NF	Noise Figure			5	dB

Video Detector (VDET)

Table 9 lists the operating parameters of video detector (VDET) **426** in accordance with a preferred embodiment of the present invention. Video is detected synchronously by applying the output of IFAMP **422** as a first input signal and

the output of VCLIM 428 as a second input signal to a double-balanced mixer (VDET 426). The output of VDET 426 is configured with a single low-pass pole set to allow 4.2 MHz to pass with a 0.5 dB loss.

TABLE 9

VDET Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
BW	Output −0.5 dB Bandwidth		4.2		MHz
V_{video}	Video Output Level		0.1		V(p—p)
NF	Noise Figure			5	dB

Video Baseband Filter (VBBF)

Table 10 lists the operating parameters of video baseband filter (VBBF) 430 in accordance with a preferred embodiment of the present invention. The detected video is band-limited to 4.2 MHz by a multiple pole transitional Gaussian response low-pass filter which has been predistorted for both compensation of SIFF 420 response and group delay characteristics conforming to FCC regulations. VBBF 430 rejects the video carrier at 20 MHz and the high level 40 MHz modulated signal at the output of VDET 426 with minimum pass-band transient distortion.

TABLE 10

VBBF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_p	Passband Corner Frequency A(ω) < 0.5 dB		4.2		MHz
f_s	Stopband Corner Frequency A(ω) > 54 dB		8.0		MHz
NF	Noise Figure			5	dB

Sync Clamp (SCMP)

Table 11 lists the operating parameters of sync clamp (SCMP) 432 in accordance with a preferred embodiment of the present invention. The sync pulses are clamped via two techniques. A diode clamp is used to define the minimum level to which the sync tip is set. Once this is achieved, after one sync pulse, a sync slicer regenerates the sync pulses only and passes these on to a phase-locked loop (PLL). The PLL acquires the horizontal sync rate and generates a sync gating pulse approximately centered in the horizontal sync pulse.

Once the PLL has achieved lock, a gated clamp circuit using the sync gating pulse is enabled to define the sync level and the diode clamp is disabled. For processing of scrambled video, a line counter is inserted into the phase-locked loop to count a field, the sync is integrated for detection of vertical sync and only the vertical sync pulse is used for the clamp gate. The sync gating pulse is used, via a clocked delay of 5 microseconds to generate a blanking gate used for AGC processing. SCMP 432 also amplifies the video to its final output level. The output signal RX LOCK 434 becomes active when the phase-locked loop has achieved lock to the sync pulses.

TABLE 11

SCMP Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		20		dB
V_{Video}	Video Output Level		1		V(p—p)
f_{PLLH}	PLL Operating Frequency H-Sync Mode		15750		kHz
f_{PLLV}	PLL Operating Frequency V-Sync Mode		60		Hz

Noise Inverter (NINV)

Table 12 lists the operating parameters of noise inverter (NINV) 436 in accordance with a preferred embodiment of the present invention. SCMP 432 video output will include extreme transient noise pulses in both black and white directions. The black noise is typically a large RF transient due to some form of interference. The signal will drop below the blanking level during the horizontal line picture period and may upset downstream sync detect circuits. A comparator detects these levels below the blanking level during active picture, or non-blanked, periods and enables an inverting video path during the event which is summed with the original video. The summation cancels the effect of the noise pulse so that the sync level region is not disturbed. Extreme positive excursions, which appear as white, are also cancelled when they exceed the equivalent 108 IRE white level, or less than 5 percent modulation. These typically are due to desensitization effects of the receiver when pulse interference exists on an out of band frequency and typically result in the disruption of the synchronous detector. Positive noise inversion is accomplished using a comparator referenced to the 108 IRE level and enabling a video inversion path to a summer as for the negative case above.

TABLE 12

NINV Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p—p)
V_{WINS}	White Noise Inverter Insertion Level		95		IRE
V_{BINS}	Black Noise Inverter Insertion Level		10		IRE
V_{WTH}	White Inverter Threshold		108		IRE
V_{BTH}	Black Inverter Threshold Non-Blanked		0		IRE

Automatic Gain Control (AGC)

Table 13 lists the operating parameters of automatic gain control (AGC) 438 in accordance with a preferred embodiment of the present invention. The clamped video is fed into a sample-and-hold circuit controlled by the blanking gate. The sample-and-hold circuit output drives a differential amplifier which has a reference level at its other input equal to the desired output blanking level for the standard NTSC video waveform.

On each horizontal sync pulse the output of the differential amplifier represents the gain error of the receiver to achieve the video waveform. This is fed back to LNTA 406 and IFAMP 422 to adjust the gain as necessary. The timing and gain of the differential amplifier must be set for fast

transient response time to account for aircraft induced flutter (fading) and unconditional stability. The response can be changed to update gain only during the vertical sync period, just as with SCMP 432, so that scrambled video may pass through the tuner without any adverse effects.

The AGC voltages defined by the performance of the RF and IF amplification behavior and input signal level, is measured via 4 bit analog-to-digital converter and stored as the data word Received Signal Strength Indicator (RSSI) 439. Automatic gain control is applied first to IFAMP 422 to reduce the video level, and beyond 50 dB of reduction, also to LNTA 406, which implements a delayed AGC function.

TABLE 13

AGC Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p—p)
$V_{s\text{-}b}$	Input Sync to Blanking Level Reference		0.286		V
G_{AGC}	AGC Loop Gain	70			dB
GBW_H	AGC Gain-Bandwidth H-Mode		150		Hz
GBW_V	AGC Gain-Bandwidth V-Mode		6		Hz
ϕ_{MAGC}	AGC Loop Phase Margin	60			°

Video Output Buffer (VBUF)

Table 14 lists the operating parameters of video buffer (VBUF) 437 in accordance with a preferred embodiment of the present invention. The video is at an internal level of 1 volt peak-to-peak differential. VBUF 437 converts this signal to a current via a linear transconductance amplifier that drives a load resistor. Feedback is arranged via an external signal to allow for continuous adjustment of the gain and offset of the amplifier so that no AC coupling is required in the next stage. An on-chip integrator is used to define the feedback level and reject the video signal. The control signal VOUT MODE 445 enables on-chip loads of 75 Ω, 1000 Ω or no load, which equates to a current source output. The control signal VIDEO MUTE 446 enables a video mute switch which sets the output to the blanking level.

TABLE 14

VBUF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p—p)
V_{OV}	Video Output Level ($R_L > 70$ Ω)		1		V(p—p)
V_{OVR}	Video Output DC Reference ($R_L > 70$Ω)	0.5		3.5	V
Z_{OUT}	Output Impedance 75Ω Mode	73	75	77	Ω
Z_{OUT}	Output Impedance 1000Ω Mode	980	1000	1020	Ω
Z_{OUT}	Output Impedance Current Mode		10		kΩ

Automatic Frequency Control (AFC) Frequency Detector

Table 15 lists the operating parameters of automatic frequency control (AFC) frequency detector (DET) 440 in accordance with a preferred embodiment of the present invention. The extracted IF video carrier is compared to a fixed reference at 20 MHz and used to drive LO1 in a direction to equalize the two. The error in LO1 from its frequency control setting is the signal AFC ERROR 452. The 20 MHz reference is generated via a frequency synthesizer from the master reference. The 20 MHz reference is compared to the extracted video carrier signal at the output of VCEF 425 using a digital frequency comparator which drives a seven bit up-down counter that is summed with the low bits of the frequency code sent to LO1 from the I²C interface 170. Thus, the frequency error appears in the counter which can be read via the interface bus. This counter is set to 1000000b when there is no frequency error. This condition sets the status signal FREQ LOCK 453.

TABLE 15

AFC Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{CIN}	Video Carrier Input Level		0.1		V(p—p)
f_{ERROR}	Frequency Error From Nominal	−500		500	kHz
f_{UPDATE}	Frequency Update Rate			30	Hz

AFC Reference Oscillator (AFCREF)

Table 16 lists the operating parameters of AFC reference oscillator (AFCREF) 442 in accordance with a preferred embodiment of the present invention. AFCREF 442 is a fixed frequency synthesizer running at 20 MHz, which is phase-locked to the quartz crystal frequency reference. Phase noise is not particularly critical due to the application of this circuit at low modulation frequencies only.

TABLE 16

AFCREF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p—p)
N_{op}	Phase Noise for 20 Hz < f_m < 1 MHz			−60	dBc
f_0	Operating Frequency		20		MHz

First Local Oscillator (LO1)

Table 17 lists the operating parameters of first local oscillator (LO1) 450 in accordance with a preferred embodiment of the present invention. LO1 is a phase-locked frequency synthesizer configured for specific levels of phase noise to allow for best receiver sensitivity without excessive spurious noise responses. It is digitally controlled and utilizes a quartz crystal reference for frequency and in-band phase noise control. The frequency control input FREQUENCY 451 from the serial digital interface 170 bus has sixteen (16) bits with 62.5 kHz LSB resolution. Four (4) more bits below the LSB as well as the first 3 LSB's are used by the AFC for fine tuning.

TABLE 17

LO1 Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p—p)
N_{op}	Phase Noise for f_m > 1 MHz			−110	dBc
N_{op}	Phase Noise for 60 Hz < f_m < 1 MHz			−80	dBc

TABLE 17-continued

LO1 Specifications

Parameter	Description	Min	Typ	Max	Unit
f_0	Operating Frequency	1200		2100	MHz
f_{STEP}	Frequency Step		3.906		kHz
f_{REF}	Reference Frequency		62.5		kHz

Second Local Oscillator (LO2)

Table 18 lists the operating parameters of second local oscillator (LO2) **412** in accordance with a preferred embodiment of the present invention. LO2 is a fixed tuned frequency synthesizer set to 1180 MHz. It uses the same frequency reference as LO1 and has similar phase noise characteristics.

TABLE 18

LO2 Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p—p)
N_{op}	Phase Noise for $f_m > 1$ MHz			−110	dBc
N_{op}	Phase Noise for 60 Hz $< f_m < 1$ MHz			−80	dBc
f_0	Operating Frequency		1180		MHz
f_{REF}	Reference Frequency		62.5		kHz

Audio IF Filter (AIFF)

Table 19 lists the operating parameters of audio IF filter (AIFF) **427** in accordance with a preferred embodiment of the present invention. The audio IF signal is tapped from the output of IFAMP **422** and run through a narrow-band filter operating a 15.5 MHz with a bandwidth of 300 kHz. AIFF **427** rejects the video carrier and the burst frequency components.

TABLE 19

AIFF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_0	Center Frequency		15.5		MHz
BW	Bandwidth		300		kHz
NF	Noise Figure			10	dB

Audio IF Limiter (AIFLIM)

Table 20 lists the operating parameters of audio IF limiter (AIFLIM) **444** in accordance with a preferred embodiment of the present invention. The audio IF is limited with a 40 dB possible variation in input level based on AGC **438** and the difference between sources of the broadcast signal. The output level accuracy over this range is better than −40 dB.

TABLE 20

AIFLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	40			dB
V_{th}	Limiter Input		2		mV (p-p)

TABLE 20-continued

AIFLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
	Threshold				
V_{limit}	Limiter Output Level		0.2		V (p-p)
NF	Noise Figure			10	dB

Audio Phase Detector (APD)

Table 21 lists the operating parameters of audio phase detector (APD) **455** in accordance with a preferred embodiment of the present invention. The audio IF is compared with APS **460** output no generate a phase error signal to drive APS **460** in a delay-locked loop. APD **455** is a phase detector configured for fast loop acquisition and minimum phase error once acquired.

TABLE 21

APD Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{IN}	Voltage Input Level		0.35		V (rms)

Audio DLL Low Pass Filter (ADLPF)

Table 22 lists the operating parameters of audio delay-locked loop low-pass filter (ADLPF) **458** in accordance with a preferred embodiment of the present invention. ADLPF **458** must not respond to any frequencies above 20 Hz in order to achieve the full audio bandwidth in the detected signal. The low-pass filter accommodates this requirement.

TABLE 22

APLPF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
BW	Loop Bandwidth			20	Hz
Z_L	Closed Loop Zero		10		Hz

Audio Phase Shifter (APS)

Table 23 lists the operating parameters of audio phase shifter (APS) **460** in accordance with a preferred embodiment of the present invention. APS **460** is a voltage controlled phase shifter which is locked to the audio carrier via a delay-locked loop. APS **460** generates a 90° degree phase shift in the output signal at the aural carrier frequency of 15.5 MHz.

TABLE 23

AQO Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{OUT}	Output Voltage		0.1		V
f_0	Center Frequency		15.5		MHz
θ_S	Phase Shift		90		Deg.

Audio Quadrature Detector (AQD)

Table 24 lists the operating parameters of audio quadrature detector (AQD) **462** in accordance with a preferred

embodiment of the present invention. AQD **462** is a mixer that is used to compare the incoming audio IF signal with a delayed version of the IF signal running in quadrature with the incoming carrier. The output is the desired audio baseband signal.

TABLE 24

AQD Specifications

Parameter	Description	Min	Typ	Max	Unit
G_v	Voltage Gain		0		dB

Audio Baseband Filter (ABF)

Table 25 lists the operating parameters of audio baseband filter (ABF) **464** in accordance with a preferred embodiment of the present invention. ABF **464** is a high-order Chebychev low-pass design with a cutoff at 120 kHz. This is wide enough to pass all multiplex audio subcarriers for external decoding.

TABLE 25

ABF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		10		dB
f_p	Passband Corner Frequency A(ω) <0.5 dB		120		kHz
f_s	Stopband Corner Frequency A(ω) >54 dB		300		kHz

Audio Output Buffer (ABUF)

Table 26 lists the operating parameters of audio output buffer (ABUF) **466** in accordance with a preferred embodiment of the present invention. The audio is at an internal level of 0.1 volts peak to peak. ABUF **466** converts this signal to a current via a linear transconductance amplifier which then drives a load resistor. Feedback is arranged via an external signal to allow for continuous adjustment of the gain and offset of the amplifier so that no AC coupling is required to the next stage. An on-chip integrator is used to define the feedback level and reject the audio signal.

The control signal AOUT MODE **467** enables on-chip loads of 600 Ω, 10 kΩ or no load, which equates to a current source output. The control signal AUDIO MUTE **468** controls an audio mute switch which disables the output.

TABLE 26

ABUF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Audio Input Level		0.1		V (p-p)
V_{OV}	Audio Output Level (R_L >500Ω)		0.35		V (p-p)
V_{OVR}	Audio Output DC Reference (R_L > 500Ω)	0.5		3.5	V
Z_{OUT}	Output Impedance 600Ω Mode	588	600	612	Ω
Z_{OUT}	Output Impedance 10 kΩ Mode	9.8	10.0	10.2	kΩ
Z_{OUT}	Output Impedance Current Mode		100		kΩ

Serial Digital Interface

Tables 27–29 define the registers, register bits and bit functions that are addressable by serial digital interface **170** in accordance with a preferred embodiment of the present invention. For example, control and status may be performed via an inter integrated Circuit (I^2) bus interface. Serial digital interface **170** contains all registers and provides access to all parallel digital on-chip functions via the serial bus.

TABLE 27

Register Definition

Register Name	Description	Register Address	Read/Write
FREQ0	Frequency - Low byte	00	Read/Write
FREQ1	Frequency - High byte	01	Read/Write
CTRL0	Functional control	02	Read/Write
CTRL1	Functional control	03	Read/Write
STAT0	Functional status	04	Read
STAT1	Functional status	05	Read
STAT2	Functional status	06	Read
CMPY0	Company code - Low byte	07	Read
CMPY1	Company code - High byte	08	Read
REV0	Revision code - Low byte	09	Read
REV1	Revision code - High byte	0A	Read

The definitions of the bits used in each register are.

TABLE 28

Register Bit Definitions

Register	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2	Bit 1	Bit 0
FREQ0	F7	F6	F5	F4	F3	F2	F1	F0
FREQ1	F15	F14	F13	F12	F11	F10	F9	F8
CTRL0	PWRDN	AMUTE	VMUTE		MODE3	MODE2	MODE1	MODE0
CTRL1					VRS1	VRS0	ARS1	ARS0
STAT0	RLOCK	FLOCK				AFC2	AFC1	AFC0
STAT1					RSSI3	RSSI2	RSSI1	RSSI0
STAT2	T7	T6	T5	T4	T3	T2	T1	T0
CMPY0	C7	C6	C5	C4	C3	C2	C1	C0
CMPY1	C15	C14	C13	C12	C11	C10	C9	C8
REV0	R7	R6	R5	R4	R3	R2	R1	R0
REV1	R15	R14	R13	R12	R11	R10	R9	R8

The status registers are read only. Attempting to write a status register will have no effect. The function of each bit is defined as:

TABLE 29

Bit Functions

Bit Name	Function	Description
F0-15	Receiver frequency	$f = 0.0625 \cdot F + k_f$ (MHz) (F15 = MSB, F0 = LSB)
PWRDN	Circuit power-down	When high the receiver is disabled, the video and audio outputs are muted and the I²C remains functional. The circuit draws minimum current.
AMUTE	Audio mute	Audio output is disabled when high.
VMUTE	Video mute	Video output is disabled when high.
MODE0-3	Receiver mode select	Selects receiver operating mode (unused, reserved for NTSC/PAL/SECAM switching).
VRS0-1	Video output impedance select	00 = 75Ω, 01 = 1 kΩ, 02 = Current Source.
ARS0-1	Audio output impedance select	00 = 600Ω, 01 = 10 kΩ, 02 = Current Source.
RLOCK	Receiver lock	High if AFC is locked to incoming signal
FLOCK	Synthesizer lock	High if frequency synthesizer locked
AFC0-2	AFC offset	Indicates magnitude and direction of mistuning of receiver. Three MSB's of AFC counter.
RSSI0-3	Received signal strength indicator	Indicates relative signal strength of signal. 0000 is no effective signal, 1111 is large signal. (RSSI3 = MSB, RSSI0 = LSB)
T0-7	Test data	Data used for test purposes only.
C0-15	Company code	Unique identifier for Cirrus Logic I²C parts
R0-15	Revision code	Unique identifier for part number and revision

Bias and Control (BC)

Table 30 lists the operating parameters of bias and control logic (BC) **180** in accordance with a preferred embodiment of the present invention. Bias voltages and currents are generated to define specific levels need for each circuit element in the present invention. The control signal POWERDOWN **476** controls the power-down function via BC **180**.

TABLE 30

BC Specifications

Parameter	Description	Min	Typ	Max	Unit
F_{OSC}	Crystal Oscillator Frequency Reference		8		MHz

Although the present invention and its advantages have been described in detail, it should be understood that various changes, substitutions and alterations can be made herein without departing from the spirit and scope of the invention as defined by the appended claims.

What is claimed is:

1. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled directly to an output of said first mixer and a second input coupled to said second reference signal, wherein said second mixer is an image rejection mixer.

2. The television receiver as set forth in claim **1** wherein said first mixer is a subtractive mixer.

3. The television receiver as set forth in claim **2** wherein a minimum frequency value of said first reference signal is greater than an input cutoff frequency of an RF signal received from said RF signal source.

4. The television receiver as set forth in claim **1** wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

5. The television receiver as set forth in claim **4** wherein said first mixer subtractively mixes said first reference signal and an RF signal received from said RF signal source to thereby generate a first IF signal on said first mixer output.

6. The television receiver as set forth in claim **5** wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

7. The television receiver as set forth in claim **6** wherein said first predetermined frequency is 1200 Megahertz.

8. The television receiver as set forth in claim **6** wherein said selected carrier frequency is selected by varying said first operating frequency.

9. The television receiver as set forth in claim **1** wherein said first mixer is coupled to said receiver input by an input filter which filters out all frequency components in said RF signal above an input cutoff frequency.

10. The television receiver as set forth in claim **9** wherein said minimum frequency value of said first reference signal is greater than said input cutoff frequency.

11. The television receiver as set forth in claim **6** wherein said second mixer mixes said second reference signal and said first IF signal to thereby generate a second IF signal containing all of the channels of information as contained in said RF signal.

12. The television receiver as set forth in claim **11** wherein said second operating frequency is fixed.

13. The television receiver as set forth in claim **12** wherein said second mixer rejects from said second IF signal an image signal of said second reference signal.

14. The television receiver as set forth in claim **13** wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

15. The television receiver as set forth in claim **13** wherein said minimum frequency value of said first reference signal is greater than an input cutoff frequency of said RF signal.

16. The television receiver as set forth in claim **1** further comprising a first IF signal generated by said first mixer wherein said second mixer mixes said second reference signal and said first IF signal to generate a second IF signal containing all of the channels of information as contained in an RF signal received from said RF signal source.

17. The television receiver as set forth in claim **16** wherein said second operating frequency is fixed.

18. The television receiver as set forth in claim **17** wherein said second mixer rejects from said second IF signal an image signal of said first IF signal.

19. The television receiver as set forth in claim **18** wherein a selected carrier frequency in an RF signal received from said RE signal source appears at a first predetermined frequency in said first IF signal and at a second predetermined frequency in said second IF signal.

20. The television receiver as set forth in claim **19** wherein said second predetermined frequency is a difference in frequency between said first predetermined frequency and said second operating frequency.

21. The television receiver as set forth in claim **20** wherein said second operating frequency is less than said first

predetermined frequency and the frequency of said image signal rejected by said second mixer is less than said second operating frequency by an amount equal to said second predetermined frequency.

22. The television receiver as set forth in claim 20 wherein said first predetermined frequency is 1200 Megahertz, said second operating frequency is 1180 Megahertz, and said second predetermined frequency is 20 Megahertz.

23. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer physically located on an integrated circuit substrate having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value; and

a second mixer physically located on the same integrated circuit substrate as the first mixer and having a first input directly coupled without leaving the substrate to an output of said first mixer and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

24. The television receiver as set forth in claim 23 wherein said minimum frequency value is greater than an input cutoff frequency of an RF signal received from said RF signal source.

25. The television receiver as set forth in claim 24 wherein said first mixer subtractively mixes said RF signal and said first reference signal to thereby produce a first IF signal, wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal and said first predetermined frequency is greater than said selected carrier frequency.

26. The television receiver as see forth in claim 25 wherein said second mixer mixes said first IF signal and said second reference signal to thereby produce a second IF signal, wherein said selected carrier frequency appears in said second IF signal at a second predetermined frequency and said second predetermined frequency is less than said first predetermined frequency and said second operating frequency.

27. The television receiver as set forth in claim 26 wherein said second mixer rejects at least one image signal associated with said second operating frequency from said second IF signal.

28. A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal, wherein said RF signal mixing and said first IF signal mixing operations are accomplished on the same integrated circuit substrate.

29. The method as set forth in claim 28 including the further step of filtering out of the RF signal all frequency components above an input cutoff frequency.

30. The method as set forth in claim 29 wherein the first operating frequency is greater than the input cutoff frequency.

31. The method as set forth in claim 30 wherein the step of mixing the RF signal with the first reference signal is a subtractive mixing of the RF signal and the first reference signal.

32. The method as set forth in claim 31 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a frequency higher than the RF signal.

33. The method as set forth in claim 28 wherein the step of mixing the first IF signal with the second reference signal further includes the substep of removing from the second IF signal at lease one image signal associated with the first IF signal.

34. A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal.

35. The method as set forth in claim 34 wherein the first operating frequency is greater than the input cutoff frequency.

36. The method as set forth in claim 35 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in the RF signal appears in the first IF signal at a first predetermined frequency and wherein the first predetermined frequency is higher than the RF signal.

37. The method as set forth in claim 34 wherein the step of mixing the first IF signal with the second reference signal further includes the substep of removing from the second IF signal at least one image signal associated with the first IF signal.

38. The method as set forth in claim 34 wherein the step of mixing the first IF signal with the second reference signal is a down conversion mixing, wherein a selected carrier signal appearing in the first IF signal at a first predetermined frequency appears in the second IF signal at a second predetermined frequency and wherein the second predetermined frequency is lower than the first IF signal.

39. The method as set forth in claim 34 including the further step of filtering the second IF signal with a bandpass filter.

40. The method as set forth in claim 39 wherein the bandpass filter has a passband between 15 Megahertz and 21 Megahertz.

41. The method as set forth in claim 39 wherein the bandpass filter is a Nyquist slope filter having a cutoff frequency of 20 Megahertz.

42. A television receiver comprising:

an input filter coupled to an RF signal source;

a first amplifier coupled to an output of said input filter;

a first mixer having a first input and a second input, wherein said first input of said first mixer is coupled to an output of said first amplifier;

a first oscillator coupled to said second input of said first mixer, wherein said first oscillator generates a first reference signal having a first operating frequency;

a second mixer having a first input and a second input, wherein said first input of said second mixer is coupled directly to an output of said first mixer without the connection of a filter therebetween;

a second oscillator coupled to said second input of said second mixer, wherein said second oscillator generates a second reference signal having a second operating frequency; and

a second filter coupled to an output of said second mixer.

43. The television receiver as set forth in claim **42** wherein said input filter is a low-pass filter having a −3 dB cutoff frequency of 900 Megahertz.

44. The television receiver as set forth in claim **42** wherein said first amplifier is a transconductance amplifier having a noise figure between 4.5 dB and 5.4 dB.

45. The television receiver as set forth in claim **42** wherein the gain of said first amplifier is automatically controlled by an automatic gain control circuit of said television receiver.

46. The television receiver as set forth in claim **42** wherein said first mixer is an up-converter which subtractively mixes a filtered and amplified RF signal having a maximum cutoff frequency with said first reference signal to thereby generate a first IF signal, wherein a lower limit of said first operating frequency is greater than said maximum cutoff frequency.

47. The television receiver as set forth in claim **42** wherein said first oscillator is a phase-locked synthesizer and said first operating frequency has an upper limit of 2100 Megahertz and a lower limit of 1200 Megahertz.

48. The television receiver as set forth in claim **42** wherein said second mixer is an image-rejection mixer which down-converts a first IF signal outputted from said first mixer to thereby generate a second IF signal.

49. The television receiver as set forth in claim **42** wherein said second oscillator is a fixed tuned-frequency synthesizer and said second operating frequency is approximately 1180 Megahertz.

50. The television receiver as set forth in claim **42** wherein said second filter is a bandpass filter having an upper passband limit of 20.75 Megahertz and a lower passband limit of 15.8 Megahertz.

51. The television receiver as set forth in claim **42** wherein said second filter is a Nyquist slope filter.

52. The television receiver as set forth in claim **51** wherein said second filter has a −6 dB gain point at a visual carrier frequency of an IF signal generated by said second mixer.

53. The television receiver as set forth in claim **52** wherein said visual carrier frequency is approximately 20 Megahertz and said second filter has a passband corner frequency of approximately 19.25 Megahertz.

54. The television receiver as set forth in claim **52** wherein said second filter is automatically tuned by measuring the level of synchronization pulses on a input of said second filter and on an output of said second filter and adjusting a cut-off frequency of said second filter so that there is a 6 dB loss on said output of said second filter.

55. The television receiver as set forth in claim **42** wherein said television receiver further comprises:

a second amplifier coupled to an output of said second filter.

56. The television receiver as set forth in claim **55** wherein the gain of said second amplifier is automatically controlled by an automatic gain control circuit of said television receiver.

57. The television receiver as set forth in claim **55** wherein said television receiver further comprises:

a third filter coupled to an output of said second amplifier, wherein said third filter is a bandpass filter for extracting a video carrier signal from an output signal of said second amplifier; and

a third mixer having a first input and a second input, wherein said first input of said third mixer is coupled to said output of said second amplifier and said second input of said third mixer is coupled to an output of said third filter and wherein an output signal on an output of said third mixer is a video baseband signal.

58. The television receiver as set forth in claim **57** wherein said second input of said third mixer is coupled to said third filter by amplitude limiting circuitry, wherein said amplitude limiting circuitry provides a fixed-amplitude reference carrier signal to said third mixer.

59. The television receiver as set forth in claim **57** wherein said television receiver further comprises frequency detection circuitry having a first input and a second input, wherein said first input of said frequency detection circuitry is coupled to said output of said third filter and said second input of said frequency detection circuitry is coupled to a third reference signal having a third operating frequency.

60. The television receiver as set forth in claim **59** wherein said frequency detection circuitry compares the frequency of said video carrier signal to said third operating frequency to thereby produce an output control signal, wherein said output control signal is used to control said first operating frequency of said first reference signal generated by said first oscillator.

61. The television receiver as set forth in claim **57** wherein said television receiver further comprises a fourth filter coupled to said output of said third mixer and wherein said fourth filter is a low-pass filter that cuts off all signal components higher in frequency than said video baseband signal.

62. The television receiver as set forth in claim **61** wherein said television receiver further includes a clamping circuit for clamping synchronization pulses in said video baseband signal from said fourth filter.

63. The television receiver as set forth in claim **62** wherein said television receiver further comprises a noise inverter circuit coupled to said clamping circuit, wherein said noise inverter circuit removes transient noise pulses from said video baseband signal.

64. The television receiver as set forth in claim **63** wherein said transient noise pulses comprise transient noise pulses that cause said video baseband signal to drop below a blanking level of said video baseband signal.

65. The television receiver as set forth in claim **63** wherein said transient noise pulses comprise transient noise pulses that cause said video baseband signal to rise above a 108 IRE white level of said video baseband signal.

66. The television receiver as set forth in claim **57** wherein said television receiver further comprises an automatic gain control circuit coupled to said output of said third mixer, wherein said automatic gain control circuit controls the gain of said second amplifier.

67. The television receiver as set forth in claim **66** wherein said automatic gain control circuit controls the gain of said first amplifier.

68. The television receiver as set forth in claim **66** wherein said automatic gain control circuit comprises a sample-and-hold circuit coupled to a first input of a differential amplifier.

69. The television receiver as set forth in claim **68** wherein a second input of said differential amplifier is coupled to a reference voltage equal to an output blanking level of a standard NTSC video baseband signal.

70. The television receiver as set forth in claim **57** wherein said television receiver further comprises a video output buffer amplifier coupled to said output of said third mixer.

71. The television receiver as set forth in claim 70 wherein said video output buffer amplifier comprises a linear transconductance amplifier.

72. The television receiver as set forth in claim 71 wherein said video output buffer amplifier further comprises blanking means for setting an output signal of said video output buffer amplifier to a video blanking level.

73. The television receiver as set forth in claim 72 wherein said video output buffer amplifier further comprises switching means for varying an output load of said video output buffer amplifier.

74. The television receiver as set forth in claim 55 wherein said television receiver further comprises a third filter coupled to an output of said second amplifier, wherein said third filter is a narrow-band filter for filtering an audio component signal from a video output signal on said output of said second amplifier.

75. The television receiver as set forth in claim 74 wherein said third filter has a center frequency of 15.5 Megahertz and a band width of 300 Kilohertz.

76. The television receiver as set forth in claim 74 wherein said television receiver further comprises:

audio phase detector circuitry coupled to an output of said third filter; and

a third mixer having a first input and a second input, wherein said first input of said third mixer is coupled to said output of said third filter and said second input of said third mixer is coupled to an output of said audio phase detector circuitry.

77. The television receiver as set forth in claim 76 wherein said audio phase detector circuitry comprises a delay-locked loop, wherein an output signal of said delay-locked loop operates in phase quadrature with a center frequency of said audio component signal on said output of said third filter.

78. The television receiver as set forth in claim 76 wherein said television receiver further comprises a fourth filter coupled to an output of said third mixer, wherein said fourth filter is a low-pass filter with a cut-off frequency of 120 Kilohertz.

79. The television receiver as set forth in claim 76 wherein said television receiver further comprises an audio output buffer amplifier coupled to an output of said third mixer.

80. The television receiver as set forth in claim 79 wherein said audio output buffer amplifier comprises a linear transconductance amplifier.

81. The television receiver as set forth in claim 79 wherein said audio output buffer amplifier further comprises muting means for setting an output signal of said audio output buffer amplifier to zero volts.

82. The television receiver as set forth in claim 80 wherein said audio output buffer amplifier further comprises switching means for varying an output load of said audio output buffer amplifier.

83. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a first bandpass filter coupled to an output of said first mixer; the bandpass filter passing more than one channel and performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent circuitry;

a second reference signal having a second operating frequency; and

an image rejection mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to said second reference signal.

84. The television receiver as set forth in claim 83 wherein said first mixer is a subtractive mixer.

85. The television receiver as set forth in claim 84 wherein a minimum frequency value of said first reference signal is greater than an input cutoff frequency of an RF signal outputted from said RF signal source.

86. The television receiver as set forth in claim 83 wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

87. The television receiver as see forth in claim 86 wherein said first mixer subtractively mixes said first reference signal and an RF signal outputted from said RF signal source to thereby generate a first IF signal on said first mixer output.

88. The television receiver as set forth in claim 87 wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

89. The television receiver as set forth in claim 88 wherein said first predetermined frequency is 1200 Megahertz.

90. The television receiver as set forth in claim 88 wherein said selected carrier frequency is selected by varying said first operating frequency.

91. The television receiver as set forth in claim 87 wherein said first mixer is coupled to said receiver input by an input filter which filters out all frequency components in said RF signal above an input cutoff frequency.

92. The television receiver as set forth in claim 91 wherein said minimum frequency value of said first reference signal is greater than said input cutoff frequency.

93. The television receiver as set forth in claim 88 wherein said second mixer mixes said second reference signal and a first filtered signal on said output of said first bandpass filter to thereby generate a second IF signal.

94. The television receiver as set forth in claim 93 wherein said second operating frequency is fixed.

95. The television receiver as set forth in claim 94 wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

96. The television receiver as set forth in claim 94 wherein said minimum frequency value of said first reference signal is greater than an input cutoff frequency of said RF signal.

97. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value;

a first bandpass filter constructed in the same substrate as said first mixer and coupled to an output of said first mixer; and

a second mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

98. The television receiver as set forth in claim 97 wherein said minimum frequency value is greater than an input cutoff frequency of an RF signal outputted from said RF signal source.

99. The television receiver as set forth in claim 98 wherein said first mixer subtractively mixes said RF signal and said first reference signal to thereby produce a first IF signal, wherein a selected carrier frequency in said RF signal

appears at a first predetermined frequency in said first IF signal and said first predetermined frequency is greater than said selected carrier frequency.

100. The television receiver as set forth in claim 99 wherein said selected carrier frequency appears at said first predetermined frequency in an output passband signal of said bandpass filter.

101. The television receiver as set forth in claim 100 wherein said second mixer mixes said output passband signal and said second reference signal to thereby produce a second IF signal, wherein said selected carrier frequency appears in said second IF signal at a second predetermined frequency and said second predetermined frequency is less than said first predetermined frequency and said second operating frequency.

102. A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a first bandpass filter to produce an output passband signal having at least four channels; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

103. The method as set forth in claim 102 including the further step of filtering out of the RF signal all frequency components above an input cutoff frequency.

104. The method as set forth in claim 103 wherein the first operating frequency is greater than the input cutoff frequency.

105. The method as set forth in claim 104 wherein the step of mixing the RF signal with the first reference signal is a subtractive mixing of the RF signal and the first reference signal.

106. The method as see forth in claim 105 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a first predetermined frequency and wherein the first predetermined frequency is higher than the selected carrier frequency.

107. The method as set forth in claim 106 wherein the first predetermined frequency appears in the output passband signal of the first bandpass filter.

108. A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a bandpass filter to produce an output passband signal having at least four channels; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

109. The method as set forth in claim 108 wherein the first operating frequency is greater than the input cutoff frequency.

110. The method as set forth in claim 109 wherein the step of mixing the amplified first filtered signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in said received RF signal appears in the first IF signal at a first predetermined frequency and wherein said first predetermined frequency is higher than the selected carrier frequency.

111. The method as set forth in claim 110 wherein the selected carrier signal appears in the output passband signal at the first predetermined frequency.

112. The method as set forth in claim 108 wherein the step of mixing the output passband signal with the second reference signal is a down conversion mixing, wherein a selected carrier signal appearing at a first predetermined frequency in the output passband signal appears in the second IF signal at a second predetermined frequency and wherein said second predetermined frequency is lower than said first predetermined frequency.

113. The method as set forth in claim 108 including the further step of filtering the second IF signal with a second bandpass filter.

114. The method as set forth in claim 113 wherein the second bandpass filter has a passband between 15 Megahertz and 21 Megahertz.

115. The method as set forth in claim 114 wherein the second bandpass filter is a Nyquist slope filter having a cutoff frequency of 20 Megahertz.

116. The method of extracting a single communications channel from a large number of such channels wherein each such channel has a defined frequency range, and wherein the large number of channels are communicated within an upper and a lower frequency boundary, the method comprising the steps of:

from an input signal containing all of the large number of channels removing all frequencies greater than the upper frequency boundary;

translating all of the channels contained within the input signal to a translated frequency range such that a desired single channel to be extracted from the input signal is positioned centrally within the translated frequency range, the translated frequency range being positioned at frequencies higher than the upper frequency boundary;

removing from the translated input signal all channels extending from the center of the translated frequency range four channels or more in each frequency direction;

translating the remaining channels within the input signal to a second translated frequency range, such that the desired single channel to be extracted from the input signal is positioned centrally within the second translated frequency range having a low end frequency boundary defined as high enough not to cause phase distortion in any subsequent filtering and a high end frequency boundary defined as a frequency beyond which a filter with proper bandwidth, noise and distortion can be integrated in a substrate; and

removal of all remaining frequencies which are not part of the desired single channel.

117. The method set forth in claim 116 wherein the last-mentioned removal step includes a receiver attenuation characteristic such that the relative amplitude of low and high frequencies within the desired single channel are corrected for vestigial sideband demodulation.

118. The method set forth in claim 116 wherein the low end and high end boundaries of the second translated frequency range are defined by the implementation of an integrated circuit continuous time filter capable of achieving single channel TV signal extraction and receiver attenuation characteristics.

119. The television receiver as set forth in claim 42 wherein said input filter limits the frequencies passed to those at or below the highest channel in a desired frequency band.

120. A television receiver comprising:

a receiver input for receiving an RF signal;

an input filter coupled to said receiver input and operating to remove all frequency components in said RF signal above an input cutoff frequency to produce a filtered RF signal;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to an output of said input filter and a second input coupled to said first reference signal;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled to an output of said first mixer and a second input coupled to said second reference signal, wherein said second mixer is an image rejection mixer.

121. The television receiver of claim 120 wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

122. The television receiver of claim 121 wherein said minimum frequency value is greater than said input cutoff frequency.

123. The television receiver of claim 120 wherein said first mixer subtractively mixes said first reference signal and said filtered RF signal to thereby generate a first IF signal at said first mixer output.

124. The television receiver of claim 120 wherein said second mixer mixes a first IF signal from said first mixer and said second reference signal to generate a second IF signal and wherein said second mixer rejects an image signal of said second reference signal from said second IF signal.

125. The television receiver of claim 120 wherein said second mixer mixes a first IF signal from said first mixer and said second reference signal to generate a second IF signal and wherein said second mixer rejects an image signal of said first IF signal from said second IF signal.

126. The television receiver of claim 120 wherein said first mixer generates a first IF signal and wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

127. The television receiver of claim 126 wherein said second mixer generates a second IF signal and wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

128. The television receiver of claim 120 wherein said first mixer and said second mixer are physically located on the same integrated circuit substrate.

* * * * *

Exh. B

EXHIBIT B – MICROTUNE CLAIM TERM DEFINITIONS

Term	Claims	Agreed Definition
"Television receiver"	83-96, 120-128	A device that receives and selects radio frequency (RF) signals within the television band, and is synonymous with "television tuner."
"Channel"	83 only	A band of frequencies allocated for carrying a television signal, for example a 6 MHz band of frequencies between 54-890 MHz.
"RF signal"	83, 120 only	Signals allocated for television, for example, from 54-890 MHz.
"Channel"	102, 108	A portion of RF spectrum allocated for transmission of a signal for a specified purpose.
"RF signal"	102, 108	See agreed term definitions, Exhibit C.
"Image" and "image signal"	124 and 125	An undesired signal that, along with the desired signal, is present at the signal input of a mixer. The image signal is located at a frequency that is twice the frequency of the desired mixer output signal, namely the intermediate frequency (IF), away from the frequency of the desired signal. The result, in the case of a basic mixer, is to create a second output signal at the same frequency as the desired output signal, thereby causing interference with the desired mixer output signal.
"Image rejection" and "rejects an image signal"	83, 124 and 125	Attenuation of an image signal to such a degree as to minimize its undesired interaction, and thereby reduce interference with the desired output signal.
"Image rejection mixer" and "image-rejection mixer"	83 and 120	A certain type of mixer that performs image rejection in addition to combining input signals. In other words, an image rejection mixer not only performs the functions of a basic mixer, but also specifically attenuates the image signal to limit interference with the desired mixer output signal.
"Subtractive mixer"	84	A mixer in which only the frequency difference signal components created by the

EXHIBIT B – MICROTUNE CLAIM TERM DEFINITIONS

		combination of the two mixer input signals are present in the output signal, the frequency sum components being removed by a filter.
"Subtractive mixing," "subtractively mixing" and "subtractively mixes"	87, 105, 108 and 123	Combining two mixer input signals in a mixer such that only the frequency difference signal components created by the mixer are present in the output signal, the frequency sum components being removed by a filter.
"Filter"	83, 91, 93, 102, 107-108, 113-115, 120	A device that passes signals of certain frequencies and attenuates signals of other frequencies.
"Filters," "filtered" and "filtering"	91, 93, 102-103, 108, 110, 113, 120 and 123	Processing a signal such that signals of certain frequencies are passed and signals of other frequencies are attenuated.
"Bandpass filter," "passband"	83, 93, 102, 107-108 and 111-115	A filter that passes signals that lie within a selected band of frequencies with low loss, called the "passband," but attenuates signals at frequencies above and below the "passband."
"Coupled to an RF signal source"	83	An electrical connection is made to a source of one or more RF signals, for example, an antenna receiving over-the-air signals or a coaxial cable for connection to a cable television system. The term "RF signal source" is not an element.
"Passing more than one channel"	83	The passband of the bandpass filter processing a television RF signal would be greater than one channel, i.e., greater than the 6-MHz bandwidth of a standard television channel in the United States.
"Having at least four channels"	102 and 108	The passband of a bandpass filter processing a television RF signal would be greater than four television channels, i.e., 24 MHz, the bandwidth of four contiguous standard television channels in the United States.
"A first bandpass filter…performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent	83	The bandpass filter accomplishes some detectable measure of image rejection, but less than all of the image rejection accomplished by the invention, and further that the bandpass filter decreases the total power content of the signal it processes, thus reducing the power-

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EXHIBIT B – MICROTUNE CLAIM TERM DEFINITIONS

circuitry"		handling requirements of circuitry that follows the bandpass filter.
"A second mixer having a first input coupled to an output of said first mixer"	120	A signal path extends from a first mixer output to a second mixer input, and there can be intermediate circuit components between the first mixer output and second mixer input.

EXC

EXHIBIT C – AGREED CLAIM TERM DEFINITIONS

Term	Claims	Agreed Definition
"Frequency"	(general); all claims with "RF", "input cutoff frequency" or "operating frequency"	The rate of periodic change of a signal. Frequencies are commonly measured in cycles per second, called Hertz, abbreviated "Hz". High frequencies are commonly measured in millions of Hertz, or "Megahertz", abbreviated "MHz;" 1 MHz is equal to 1,000,000 Hz.
"RF Signal"	102 and 108	A signal containing one or more frequencies in the radio frequency (RF) domain, i.e., from 10,000 Hz to 300,000 MHz. Note: dispute on whether limited to TV band in claims with "television" in preamble
First bandpass filter "constructed in the same substrate" as said first mixer	97	The first bandpass filter and the mixer are both located in the same integrated circuit chip.
Mixer(s) . . . "physically located on the same integrated circuit substrate"	23 and 128	The mixer(s) are located internally in one integrated circuit (IC) chip.
Second mixer "coupled directly" to first	1 and 42	The second mixer is directly connected to the first mixer with no other circuit components between the two mixers.
"Nyquist slope filter"	41, 51 and 115	A filter applied to a VSB modulated signal just prior to detection to restore flat frequency response.
"Mixer"	1, 23, 42, 83, 97 and 120	A device which receives electrical signals at two functional inputs and outputs an electrical signal at one functional output. The output electrical signal has components at the sums of the input frequencies and the differences of the input frequencies, and other components at frequencies related to the input frequencies.
"Mixing"	28, 31, 32, 33, 34, 36, 37, 38, 102, 108, 110 and 112	Receiving electrical signals at two functional inputs and outputting an electrical signal at one functional output. The output electrical signal has components at the sums of the input frequencies and the differences of the input

EXHIBIT C – AGREED CLAIM TERM DEFINITIONS

		frequencies, and other components at frequencies related to the input frequencies.
"Prior to eliminating any channels"	28 and 34	No filtering or significant attenuation of channels contained within the output of the first mixer is performed prior to the second mixer.
"To be processed by subsequent circuitry"	83	The circuit components immediately following the first bandpass filter. Note: dispute on whether to construe under 35 USC 112, 6th paragraph.
"For receiving an RF signal"	120	Connection to a source of RF signals, such as an antenna or cable system (a source of RF signals is not required by claim 120).
"Amplifying"	34 and 108	Increasing the voltage or current of a signal in a specified manner, causing the overall signal power to be increased.
"Directly coupled without leaving the substrate"	23	There is a direct connection in the integrated circuit chip between the output of the first mixer and the first input of the second mixer.

IN THE UNITED STATES DISTRICT COURT
EASTERN DISTRICT OF TEXAS
SHERMAN DIVISION

MICROTUNE (TEXAS), L.P.,	§	
Plaintiff	§	
	§	
v.	§	CASE NO. 4:01CV23
	§	
BROADCOM CORPORATION,	§	
	§	
Defendant.	§	
	§	

DECLARATION OF BERNARD J. LECHNER

I, Bernard J. Lechner, declare as follows:

1. Attached to this declaration are the following exhibits:

INDEX OF ATTACHED EXHIBITS	
Exhibit 1	U.S. Patent No. 5,737,035 (Rotzoll)
Exhibit 2	U.S. Patent Application Serial No. 08/426,080, specification and drawings as filed (Paper no. 1);
Exhibit 3	Office Action dated February 9, 1996 (Paper no. 4)
Exhibit 4	Amendment dated May 8, 1996 (Paper no. 5)
Exhibit 5	Amendment dated October 1, 1996 (Paper no. 7)
Exhibit 6	Amendment dated March 12, 1997 (Paper no. 9)
Exhibit 7	Amendment dated July 2, 1997 (Paper no. 11)
Exhibit 8	Article entitled "A CMOS 1.8GHz Low-Phase-Noise Voltage-Controlled Oscillator with Prescaler"
Exhibit 9	Article entitled "Future Directions in Silicon ICs for RF Personal Communications"
Exhibit 10	Article entitled "Noise in Distributed MESFET Preamplifiers"
Exhibit 11	U.S. Patent No. 4,162,452 (Ash)
Exhibit 12	U.S. Patent No. 4,340,975 (Onishi et al.)
Exhibit 13	Biography of Bernard J. Lechner

I. OVERVIEW OF DECLARATION

2. This declaration is broken into six main parts, including this overview. The other parts are my background (Part II), a tutorial (part III), an overview of the '035 patent and the described technology (Part IV), an explanation how of the invention operates (Part V) and claim construction (Part VI).

3. In order to prepare this declaration I reviewed U.S. Patent No. 5,737,035 (attached as Exhibit 1) and its file history. Exhibits 2-7 are copies of documents from the file history.

4. I also reviewed certain other documents that are attached. Exhibits 8-10 are copies of articles cited elsewhere in this declaration (Part V). Exhibits 11-12 are copies of U.S. Patents cited elsewhere in this declaration (Part V).

5. Exhibit 13 is a copy of my most recent biography.

II. BACKGROUND AND EXPERIENCE

6. I currently reside at 59 Carson Road, in Princeton, N.J. I have been retained by Microtune (Texas), L.P., ("Microtune") as an expert in the above-captioned litigation. I am over the age of eighteen and I am a citizen of the United States.

7. I have been actively involved in the field of television, from a technical perspective, since 1947. My professional engineering career began at RCA Laboratories in 1957, and continues to the present day. I am currently involved with various committees and organizations devoted to the advancement of television technology.

8. From 1957 to 1987 I worked at RCA Laboratories in the area of research and development. My work at RCA involved all aspects of television technology, including the broadcast and reception of television signals.

9. I received two RCA Laboratories Outstanding Achievement Awards and a David Sarnoff Team award in Science for my work at RCA.

10. I received the BSEE degree from Columbia University in 1957.

11. I chaired the Advisory Commission for Electrical Engineering at Mercer County Community College for 15 years. I have been a member of the Board of Directors of Palisades Institute for Research Services since 1981.

12. I am a Fellow of the Institute of Electrical and Electronics Engineers (IEEE), the Society for Information Display (SID) and the Society of Motion Picture and Television Engineers (SMPTE). I have served as Treasurer, Secretary, Vice President and President of SID. I am a member of three engineering and science honors organizations: Tau Beta Pi; Eta Kappa Nu; and Sigma Xi.

13. I am an active participant in professional organizations in the televisions area. I participated in the work of the FCC Advisory Committee on Advanced Television Service, and the Advanced Television Systems Committee (ATSC). I am chairman of the ATSC Specialists Group on Digital Television Transport Standards and I chaired the CEA/NCTA Digital Standards Subcommittee relating to standards for "Cable-Ready" Digital Television receivers.

14. I have received awards from my contributions to television technology. I received the David Sarnoff Gold Medal from SMPTE in 1996 for my contributions to television technology. In May of 2000 I was honored as the first recipient of the ATSC Outstanding Contributor Award, which has become an annual award named in my honor. In November 2001 I received the SMPTE Progress Medal Award. The Progress Medal is SMPTE's highest honor and was presented to me in recognition of my outstanding technical contributions to the progress of the engineering phases of the television industry.

15. I hold ten U.S. Patents in the area of television systems and displays.

16. Attached hereto as Exhibit 13 is a copy of my most recent biography describing my education, employment history, awards and my participation in professional organizations.

17. I have reviewed the Microtune Patent (U.S. Patent No. 5,737,035, hereinafter the '035 patent), which concerns a television tuner. I have also reviewed the file history for the '035 patent.

18. In order to understand the '035 Patent, some background information on television and television signals should be understood.

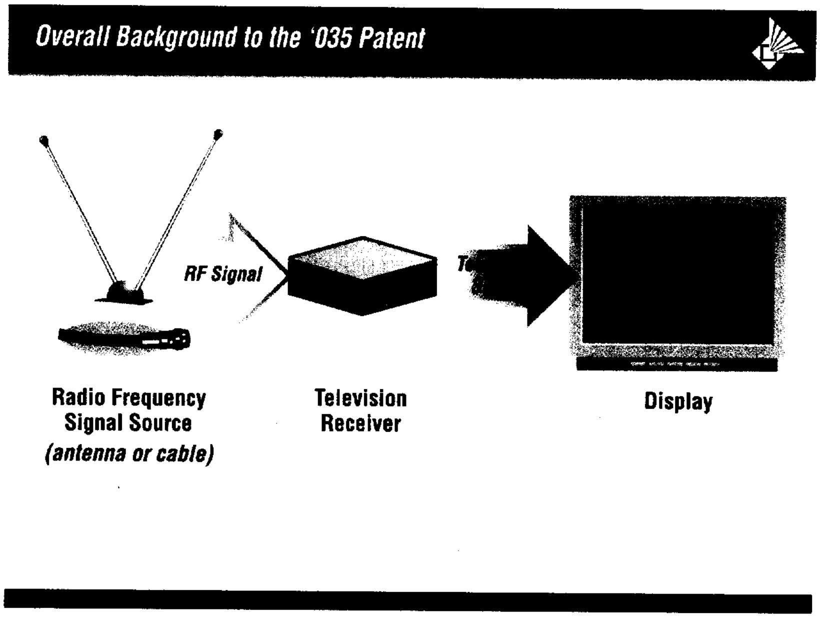
III. TUTORIAL – BACKGROUND

19. In the U.S. and abroad, television programs originate from video and audio signals captured by television cameras and audio systems. The video and audio signals are then combined to form what is called a television signal, which in turn is modulated onto a carrier signal in the radio frequency (RF) range (discussed below). RF signals have properties suitable for long-range transmission and, by using a plurality of different carriers, many different television signals can be transmitted to viewers simultaneously. The RF signals may be sent to viewers by a broadcaster over-the-air from a transmitting antenna or sent by a cable television system operator over a cable. Each separate television signal having its own distinct carrier frequency is called a television channel.

20. Although the television transmission system was originally developed to transmit pictures and sound to viewers, it is equally capable of transmitting digital information. An early example of this capability, first introduced in the United Kingdom in the 1970's by the BBC is Teletext. Teletext uses a portion of the television signal to transmit digital signals that represent pages of text and graphics for display on a television screen. Subsequent developments, both in the United Kingdom and the United States, led to systems where the entire television signal was devoted to the transmission of digital data instead of pictures and sound. These systems can be considered as predecessors to today's use of television channels for communicating digital data to computers.

21. A typical viewer connects the RF input on a television set to a receiving antenna or to a cable television system to obtain the RF signal. The viewer's television set contains a device, called a "television tuner" or "television receiver," which allows the user to extract a single channel from the RF signal. This is shown in Graphic Exhibit 1, below.

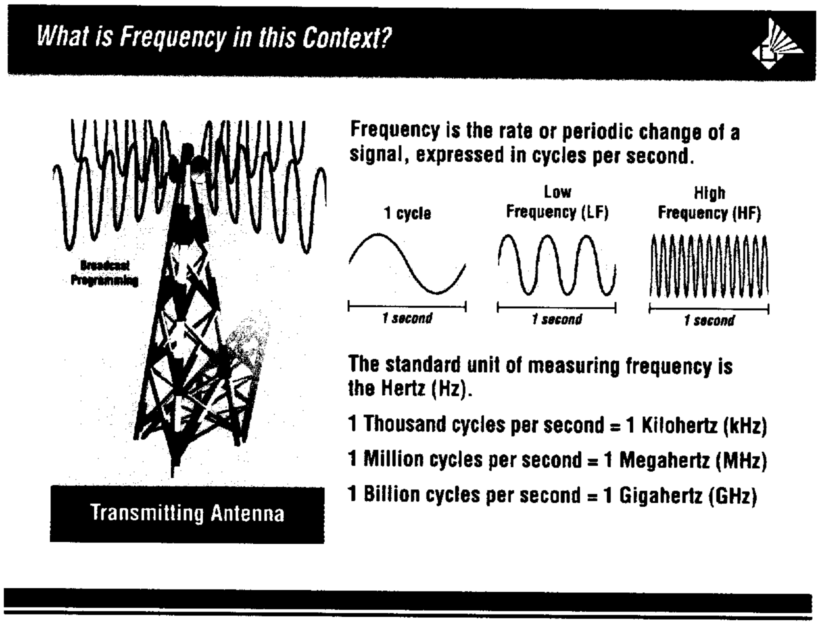
Graphic Exhibit #1



22. While Graphic Exhibit 1, above, shows a "television receiver" and "display" separately to emphasize the different functions of each, a typical television set has both a "display" and "television receiver" included within the single television set.

23. As I stated above, television signals are carried to the user by sending radio-frequency (RF) signals over-the-air, or in another medium, such as a cable. To understand RF, I will first discuss the concept of "frequency." Graphic Exhibit 2, below shows what is meant by the term "frequency."

Graphic Exhibit #2



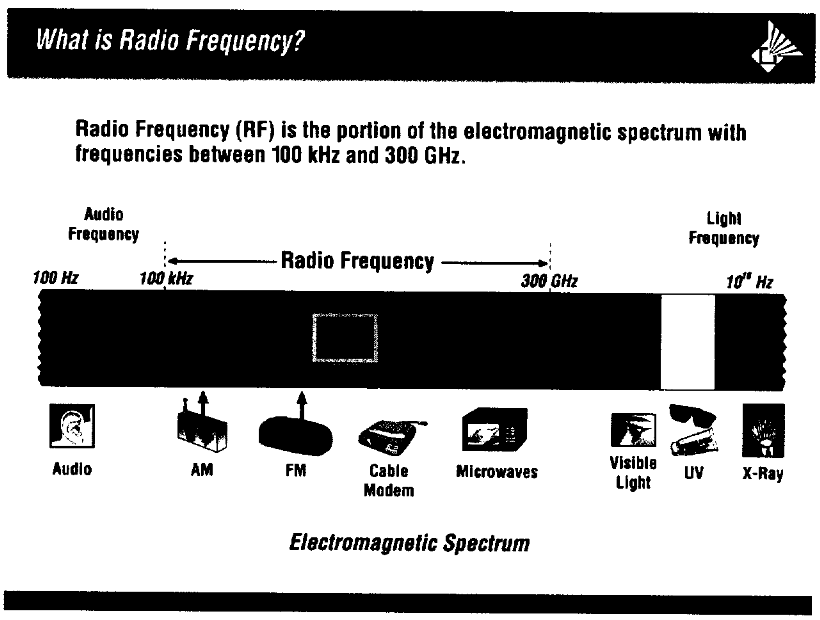
24. As shown in Graphic Exhibit 2, "frequency" relates to the number of occurrences of a periodic event per unit of time. The events can be any type of periodic event, such as ocean waves arriving on a beach, vibrations of a piano string or electrical signals including radio frequency (RF) signals. With periodic signals, such as those shown in Graphic Exhibit 2, the "frequency" is the number of times the signal repeats over a given amount of time, usually expressed as the number of repetitions (or cycles) per second. For example, a "low" frequency, shown in the second box of Graphic Exhibit 2, would be an electronic signal where three complete cycles or repetitions occur in one second. A higher frequency, shown in the third box of Graphic Exhibit 2, would be an electronic signal where 13 complete cycles occur in one second. As shown in Graphic Exhibit 2, high frequency signals have more cycles per unit of time than low frequency signals.

25. Frequencies are measured in units called "Hertz," (abbreviated Hz) which is simply the number of periodic events that occur per second. For example, the power going into a residential home has 60 complete cycles per second during each of which the voltage goes from zero to its maximum positive value, returns to zero, then goes to its maximum negative value,

and finally returns again to zero. The power line frequency is thus 60 cycles per second or 60 Hz. Because electronic signals often occur at relatively high frequencies, conventions have been developed to refer to higher frequency signals by using the term "Kilohertz" (abbreviated kHz), "Megahertz" (abbreviated MHz) and "Gigahertz" (abbreviated GHz). One Kilohertz is one thousand Hertz; one Megahertz is one million Hertz; and one Gigahertz is one billion Hertz. So 1 kHz is 1,000 Hz, 1 MHz = 1,000,000 Hz and 1 GHz = 1,000 MHz, as shown in Graphical Exhibit 2.

26. The terms "radio frequency signals" or "RF" signals mean electrical signals that fall in the range from about 100,000 Hz (or 100 kHz) to about 300,000 MHz (or 300 GHz). This is shown in Graphic Exhibit 3. The range of RF frequencies is sometimes referred to as the "RF spectrum" or "RF band of frequencies."

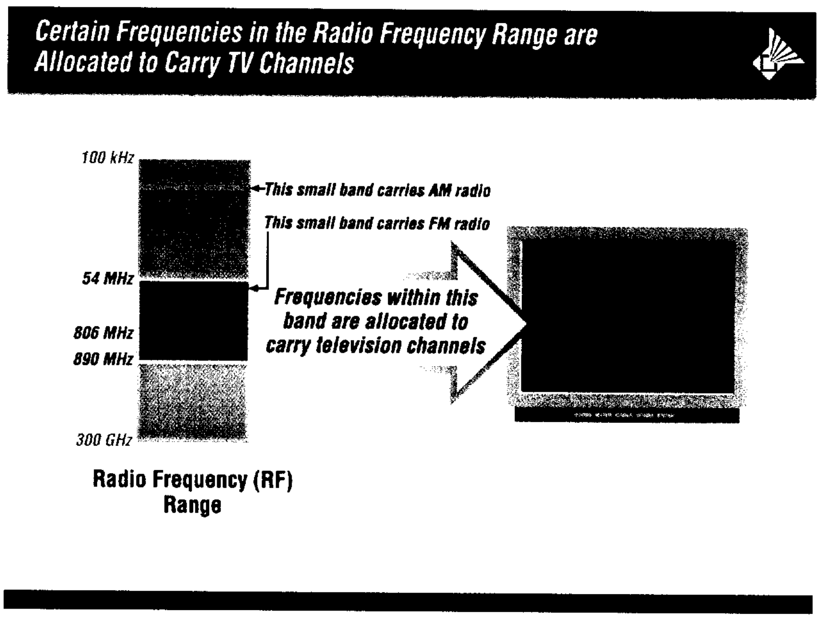
Graphic Exhibit #3



27. Within the broad RF range, or "RF spectrum" many different types of RF signals are present. For example, signals relating to television, FM radio, cellular telephones, pagers and most radar systems all occur between 30 MHz and 3 GHz. For broadcast television, over-the-air

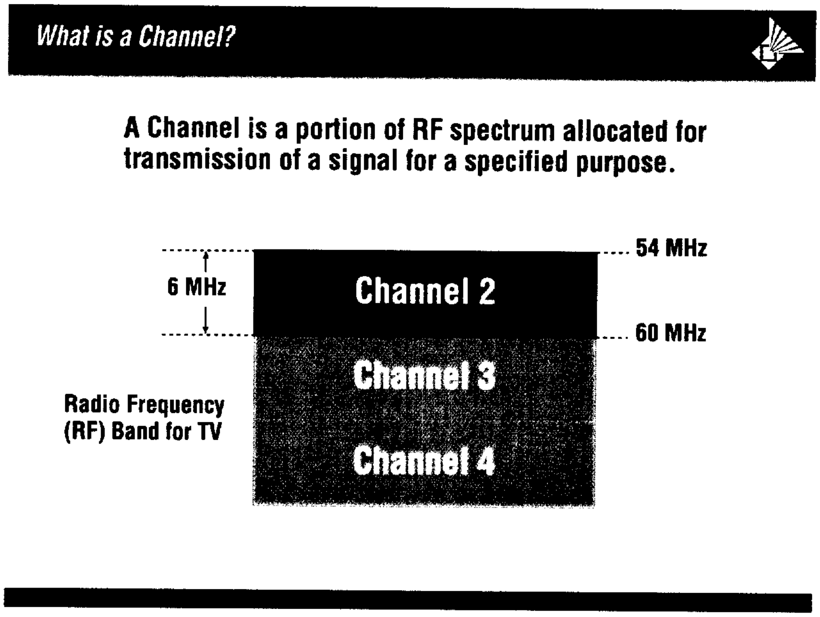
television signals in the United States are present in non-contiguous bands between about 54 MHz and 890 MHz. *See* Ex. 1, at col. 8, ll. 40-51. This range is sometimes called the "broadcast television spectrum" or "broadcast television band of frequencies." The FCC reallocated part of this spectrum, from 806 to 890 MHz (channels 70 to 83), to land mobile radio use in 1975, and its reception by television receivers was made voluntary in July 1982. Cable television has a slightly different spectrum, which is currently defined to run from 54 MHz to 864 MHz. Graphic Exhibit 4, below, shows that television occupies only a small part of the RF spectrum. Also shown in Graphic Exhibit 4, is the FM band of frequencies for FM radio broadcasts. The FM band is allocated between 88 MHz and 108 MHz, and occupies much less space than the television band.

Graphic Exhibit #4



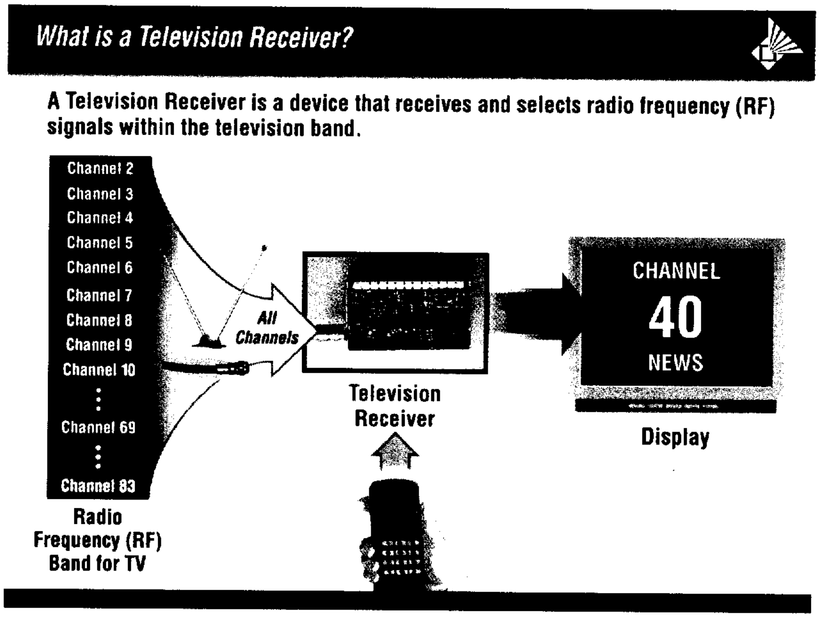
28. Within the television band of frequencies, many different television signals or "channels" are present. Each channel occupies its own band of frequencies, for which 6 MHz of bandwidth is allocated in the United States. Each channel is capable of carrying video, audio, and/or data. *Id.* at col. 8, at ll. 41-51. This is shown in Graphic Exhibit 5, below.

Graphic Exhibit #5



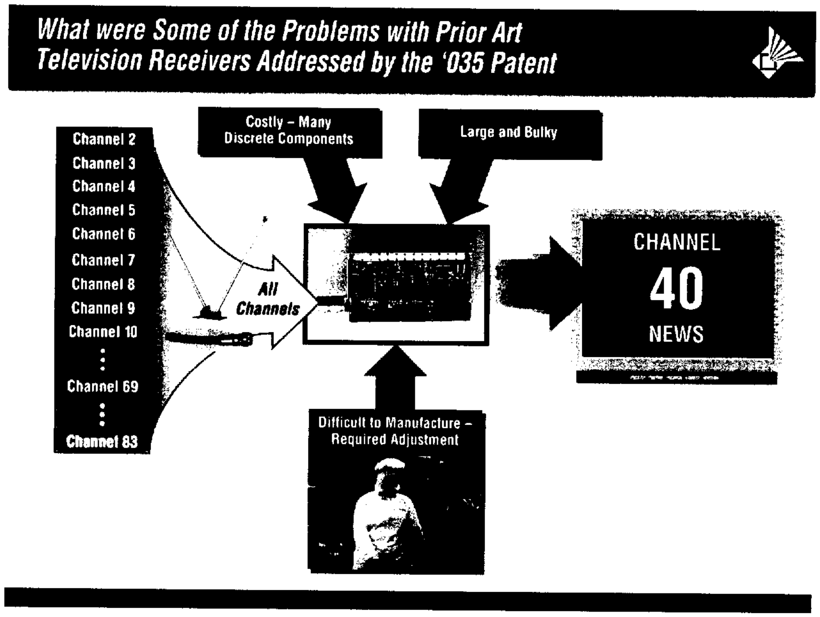
29. Since a television set is typically presented, at its input, with many television channels within the band of television frequencies, a device is needed to select a channel for viewing or, in the case of a channel carrying only digital data, for processing by a computer. For example, a viewer wanting to watch news on channel 4 must be provided with a device to select channel 4. Similarly, a computer expecting data transmitted on channel 17 must be provided with a device to select channel 17. The device that selects and extracts a channel from the television band is called a "television receiver" or "television tuner." Graphic Exhibit 6 provides an illustration of a television receiver.

Graphic Exhibit #6



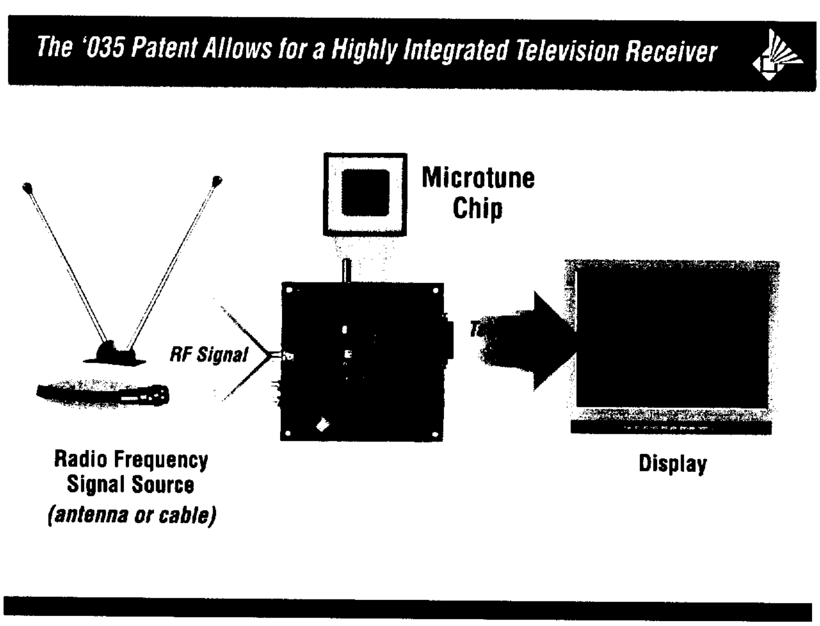
30. Conventional television tuners, such as the one shown in Graphic Exhibit 6, contain many individual circuit components, and are difficult to manufacture because they require adjustment as part of the manufacturing process. While a conventional tuner is not extremely large, it is too large and bulky to use in many electronic devices that can now be configured to receive television signals, such as laptop computers. Thus a further disadvantage of conventional tuners is their size. The disadvantages of conventional tuners are illustrated in Graphic Exhibit 7.

Graphic Exhibit #7



31. The '035 patent is directed to making a fully or partially integrated television receiver, i.e., most functions of the receiver are integrated into a single integrated circuit, or chip. This is shown in Graphic Exhibit 8, below. A chip-based receiver minimizes the number of individual circuit components, minimizes or eliminates adjustments, and decreases size and cost.

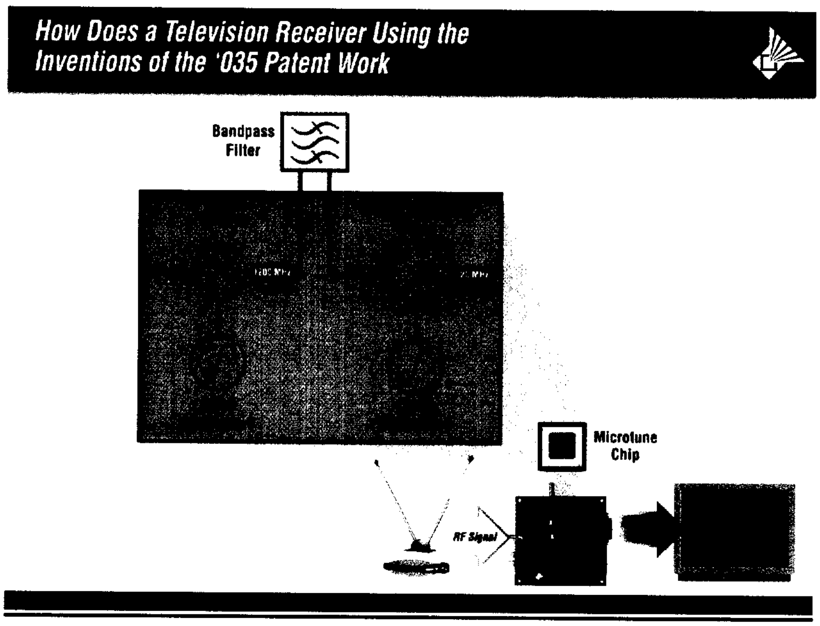
Graphic Exhibit #8



IV. OVERVIEW OF THE '035 PATENT AND DESCRIBED TECHNOLOGY

32. The relevant details of the '035 patent concern the connection of a television receiver to a source of television signals (such as an antenna or cable television system), and set out various circuit components and steps to mix, filter and further mix the television signal to select a single channel for viewing. These steps are shown schematically in Graphic Exhibit 9, below.

Graphic Exhibit #9



33. In general, the function of a television receiver is to select one channel that is to be received from the full television RF spectrum it receives. Basically, a television receiver accomplishes the reverse of the television transmission process, in which a television signal is modulated onto a carrier signal at a specific frequency to create a distinct television channel. The function of a television receiver is to select the particular frequency range allocated for the desired channel, convert it to a predetermined intermediate frequency (IF) to facilitate the rejection of adjacent channels and thereby reduce interference, and ultimately to extract the

original information that was transmitted, whether that be video and audio signals or data transmitted over a television channel, as in the case of a cable modem.

34. In order to perform the frequency conversion described in paragraph 33 above, the television receiver of the '035 invention utilizes a device called a "mixer." A "mixer" has two inputs and one output. A basic mixer "mixes" the signals on the two inputs to provide an output signal that is related to the two input signals in a way that is governed by a mathematical equation. Specifically the output signal is the mathematical product of the two input signals.

35. The output signal of a mixer contains components whose frequencies are the sum and difference of the frequencies of the input signals. The frequencies of the sum component at the mixer output are obtained by adding the frequencies of the two signals present at the input.

36. The frequencies of the difference component at the mixer output are obtained by subtracting the frequencies of one mixer input from the frequencies of the other mixer input. For purposes relevant to this case, the lower frequency signal is subtracted from the higher frequency signal to determine the resulting mixer output frequencies.[1]

37. A mixer output signal also contains other components, including unwanted distortion components, that occur at other frequencies related to the input frequencies. To summarize, a mixer output signal contains the sum and difference components of the input signals, as well as distortion components related to the input signals. This is shown in the following chart.

Mixer Output signal contains frequency components from the combination of:	
1. Frequencies on first input + frequencies on second input	(sum component)
2. Frequencies on first input – frequencies on second input	(difference component)

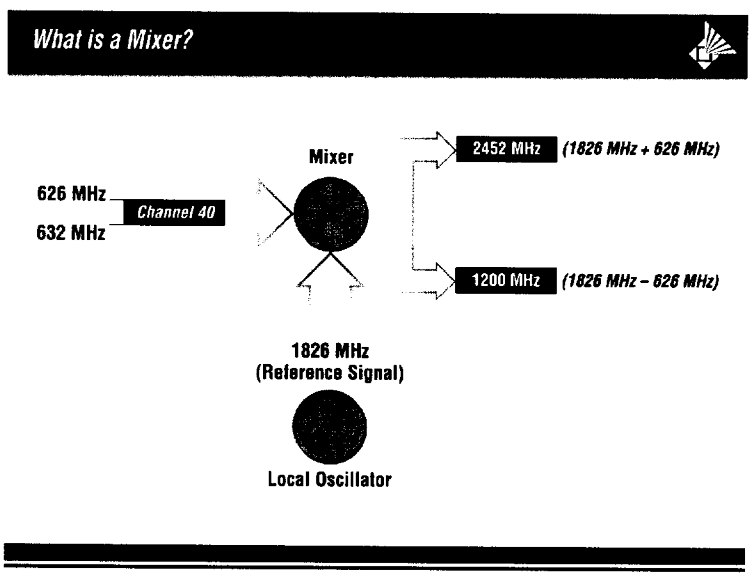
[1] An alternative method to determine the mixer output signal frequencies is to subtract the frequencies present within one mixer input signal from the frequencies present within the other mixer input signal, and convert any resulting negative frequencies to positive frequencies.

38. For example, if one mixer input is a signal with a frequency of 626 MHz and the other input is a signal with a frequency of 1826 MHz, the output will be as follows:

Mixer Output signal contains frequency components from the combination of:	**Component**	
1. Frequencies on first input + frequencies on second input	1826 + 626 = **2452** MHz	(sum component)
2. Frequencies on first input – frequencies on second input	1826 – 626 = **1200** MHz	(difference component)
Combine the Frequency components:	**Mixer output is signals at frequencies of 2452 MHz and 1200 MHz, with other distortion components.**	

39. Graphic Exhibit 10, below, shows this example in schematic form. Note that a "mixer" is denoted by a circle with an "x" in the middle. In this example, one input is labeled "626 – 632 MHz" and the other input is labeled "1826 MHz (Reference Signal)." The 1826 MHz signal is shown coming from a "local oscillator," depicted as a circle with a wave inside it.

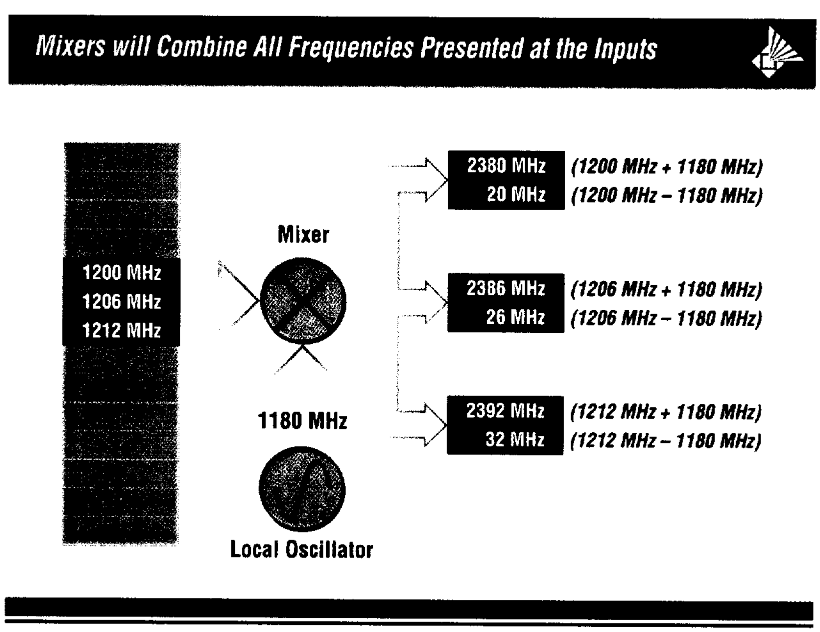
<u>**Graphic Exhibit #10**</u>



40. A "local oscillator" is a standard device, well known in the television art, which generates electrical signals of a certain desired frequency. A local oscillator may be "fixed," in which case it generates a signal at one frequency only, or "variable," in which case it can generate a signal at any frequency within a band of frequencies. While a local oscillator is intended to generate only one frequency at a time, local oscillators are not perfect and will usually generate other unwanted frequencies, including multiples of the intended frequency, called "harmonics." By using a local oscillator as one of the inputs to the mixer, the frequencies of the signals present at the output of the mixer can be varied in direct relation to the change in frequency of the local oscillator input signal.

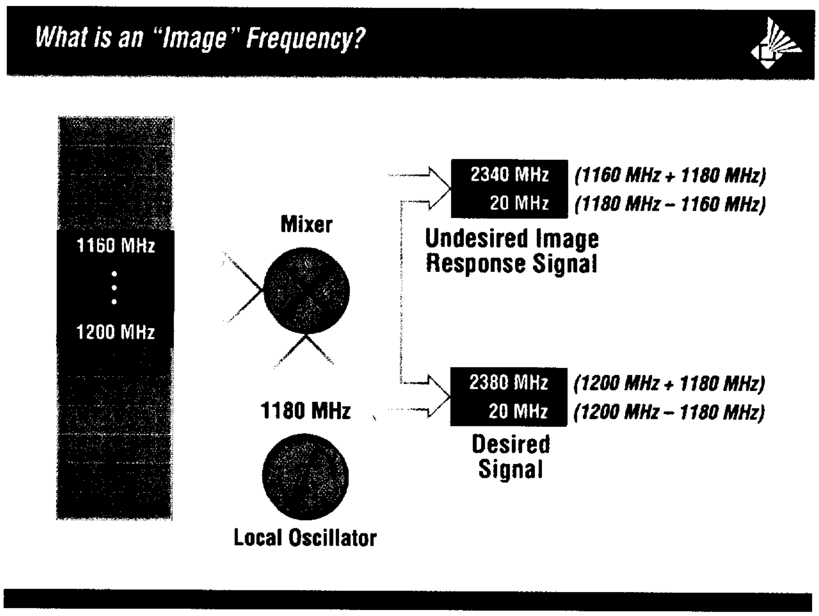
41. One characteristic of a mixer is that, if more than one signal is present at an input, the mixer will produce output signals for all of the signals present at its inputs according to the rules above. Thus if signals that span a "range" of frequencies are presented at its input, a mixer will create signals at its output that also span a "range" of frequencies. This is shown in Graphics Exhibit 11, below.

Graphic Exhibit #11



42. Because a mixer operates on a range of input frequencies (for example, television channels 2 through 83, as depicted), if a range of input signals is presented to one of a mixer's inputs, there exists the possibility of an undesired mixer output, called an "image" response. An "image" is shown below in Graphic Exhibit 12.

<u>Graphic Exhibit #12</u>

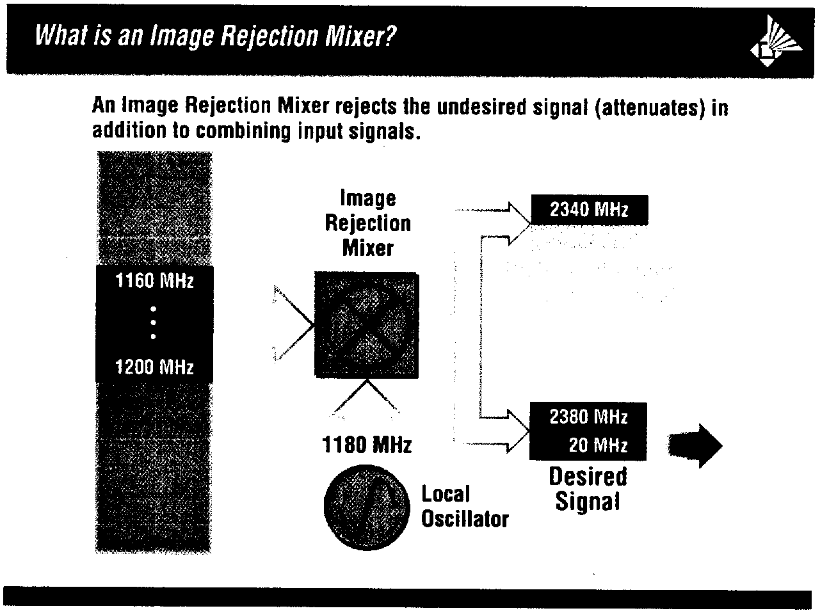


43. As shown in Graphic Exhibit 12, signals spanning a range of frequencies, including the spectrum from 1160 MHz to 1200 MHz, are fed into one mixer input. The other mixer input is at 1180 MHz. It is desired to obtain the difference component at 1200 MHz – 1180 MHz = 20 MHz.

44. However, the mixer will produce an undesired signal, also at 20 MHz, from the difference signal created when the image signal frequency of 1160 MHz is subtracted from the frequency of the local oscillator signal at 1180 MHz. This is also shown in Graphic Exhibit 12, above.

45. A special type of mixer, called an "image rejection mixer" is designed to attenuate (which means reduce) the undesired image signal. An image rejection mixer is shown below in Graphic Exhibit 13.

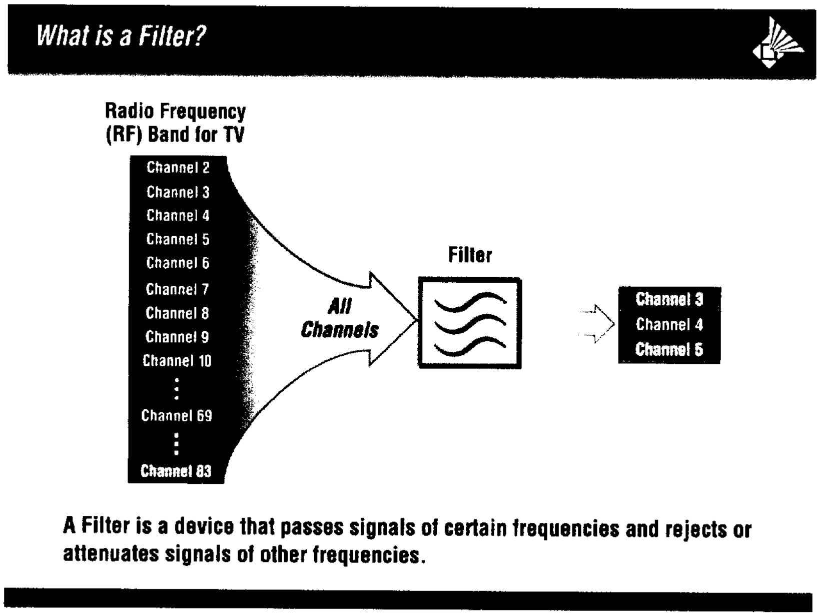
Graphic Exhibit #13



46. As shown in Graphic Exhibit 13, above, an image rejection mixer is symbolized by a circle with an "x" in the middle, located inside a square. This is shown in Figure 4 of the '035 patent. An image rejection mixer is specially constructed to attenuate the undesired image signal, as shown in Graphic Exhibit 13, above.

47. Another device disclosed in the '035 patent and known in the industry is a "filter." A filter is shown below in Graphic Exhibit 14. In the context of a television tuner, filters are used to eliminate the undesired channels and allow the selected channel to pass through.

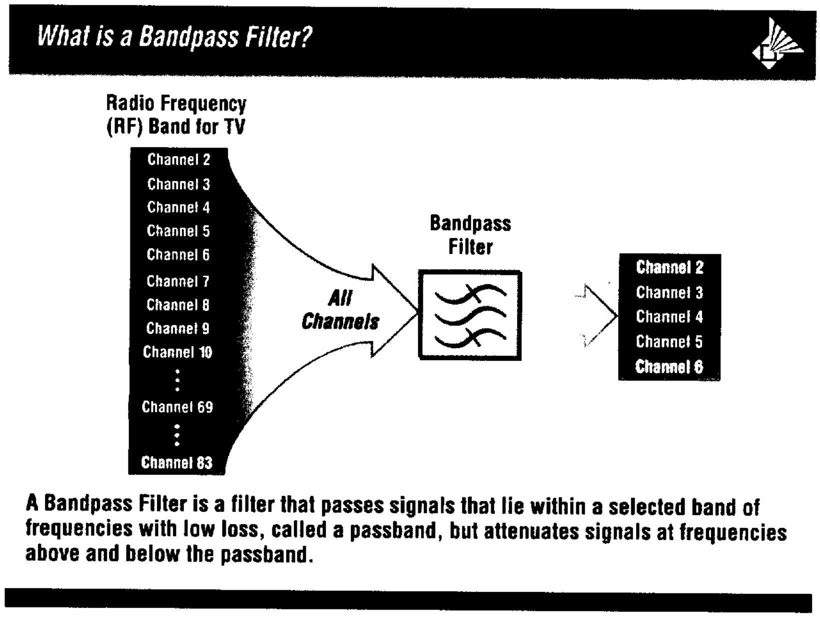
Graphic Exhibit #14



48. A filter is designed to permit only signals at certain frequencies to pass through while attenuating signals at all other frequencies. For example, a filter could be designed to pass only channel 4 and attenuate adjacent channels 2-3 and 5-83, as shown in Graphic Exhibit 14, above. One use for a filter in a television tuner is to eliminate (filter out) the undesired image signal at the input to a mixer.

49. There are different types of filters, for example, a "bandpass filter" is a type of filter that passes signals that lie within a selected band of frequencies, called the "passband" and attenuates signals at frequencies above and below the passband. Graphic Exhibit 15 shows a bandpass filter that passes three channels (in this case channels 3-5). By using an image rejection mixer, the specification for such a bandpass filter can be relaxed and the cost reduced because the bandpass filter and image rejection mixer each provide a level of attenuation of the image frequency.

Graphic Exhibit #15

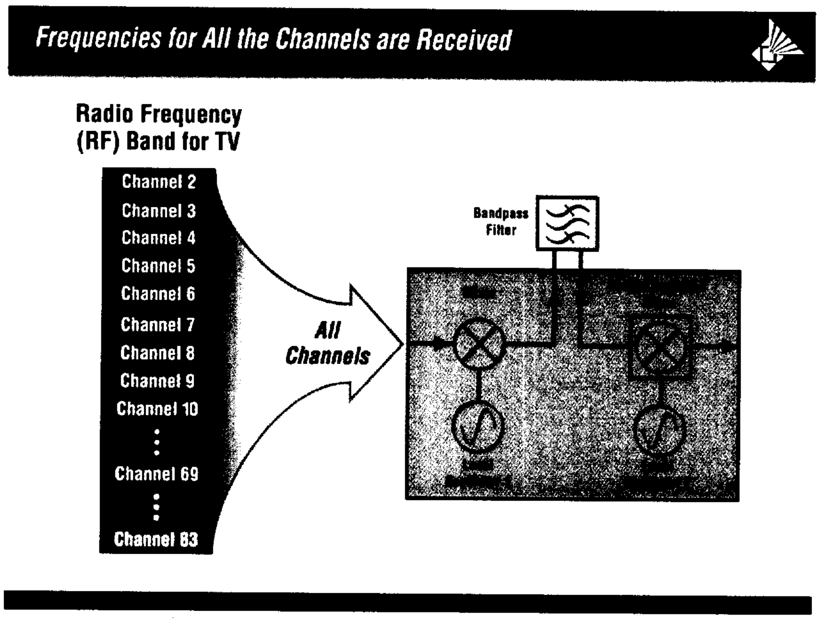


V. OPERATION OF THE INVENTION.

50. I will give an example, in general, of how a device built according to the invention in the '035 patent operates. This discussion is for overview purposes only, and includes elements present in some claims but absent from other claims, and does not address every element of every claim. The example provides an overview of the more difficult concepts of the invention, and traces the path showing how a signal is received, mixed, filtered and mixed again. The following example will trace the operation of the invention as channel 40 is selected and tuned.

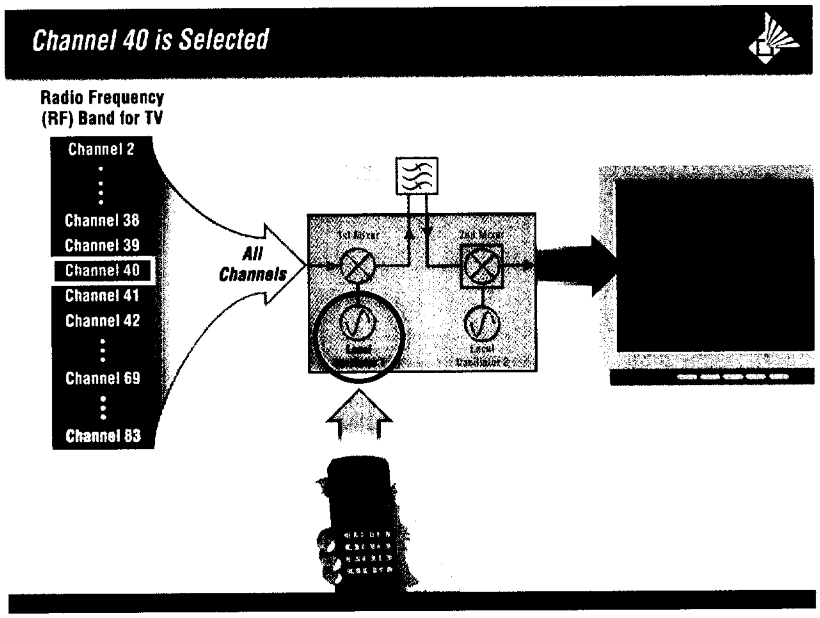
51. The input to the television tuner is typically the entire television RF band, namely from 54 to 890 MHz. In some embodiments of the invention, the RF signal is filtered by a low-pass filter to prevent signals at frequencies greater than 900 MHz from entering the tuner. Graphic Exhibit 16 illustrates all of the channels being received by the tuner.

Graphic Exhibit #16



52. In this example, the viewer desires to view channel 40. The viewer selects channel 40 by using a remote control as shown in Graphics Exhibit 17. The selection of channel 40 on the remote control causes the first local oscillator to generate a reference signal that corresponds to channel 40. The resulting output of the first mixer is the frequency band associated with channel 40, which has been converted to the desired first mixer output at 1200 MHz, known as the first intermediate frequency (IF). In this same manner, the selection of any arbitrary channel results in the selected channel being frequency-converted to the same 1200 MHz first IF. It is advantageous to convert all channels to one predetermined frequency, because it permits the use of fixed filters, such as the bandpass filter between the first and second mixers, which can be constructed with greater accuracy and lower cost than filters that must themselves be tuned across some range of frequencies.

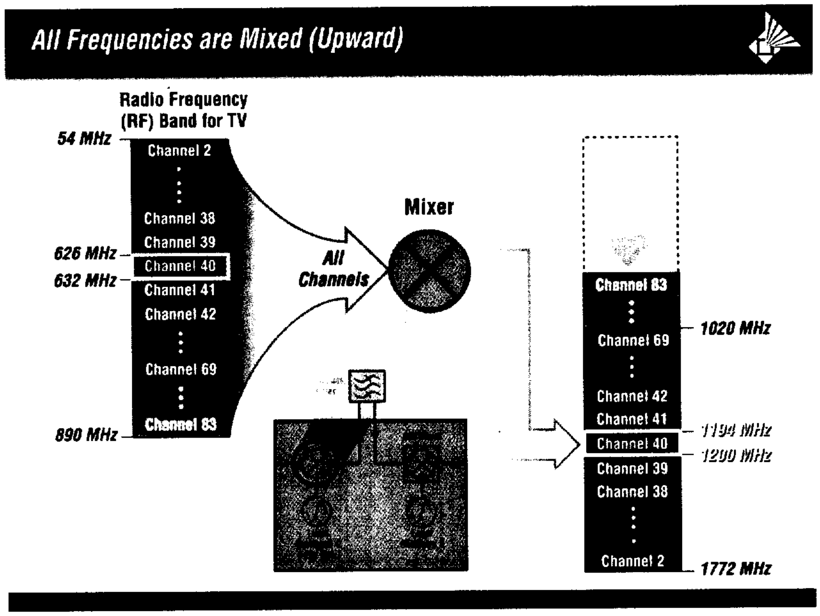
Graphic Exhibit #17



53. Channel 40 in the United States is located in the UHF band between 626 and 632 MHz. In order to cause this channel to be converted to a first IF of 1200 MHz, it is necessary to tune the first local oscillator to 1826 MHz[2]. This will result in the first mixer creating two components: a difference component at 1200 MHz and a sum component at 2452 MHz, the latter not being utilized. A side effect of using the difference component is that the channel is inverted, such that the base of the channel in the RF spectrum (626 MHz) is now the top of the channel at 1200 MHz, and conversely the top of the channel (632 MHz) in the RF spectrum is now converted to 1194 MHz, as shown in Graphic Exhibit 18 below. As stated above in the preceding paragraph, all channels will be selectively tuned such that the desired channel is always converted to the first IF of 1200 MHz.

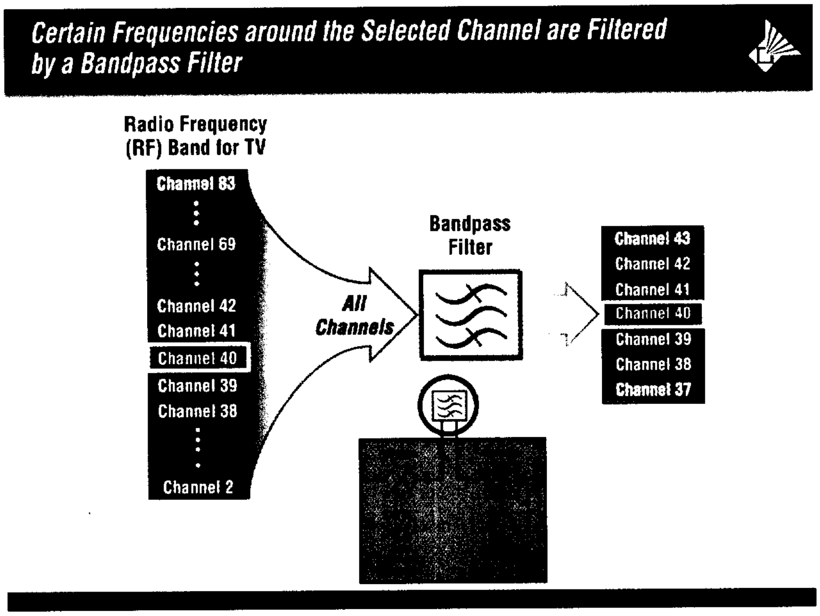
[2] Actually, since the video carrier frequency for channel 40 is located at 627.25 MHz, the local oscillator frequency must be 1827.25 MHz to produce a first IF carrier frequency of 1200 MHz. For simplicity in illustrating the operation of the invention of the '035 patent, we will ignore this 1.25 MHz offset and use round numbers.

Graphic Exhibit #18



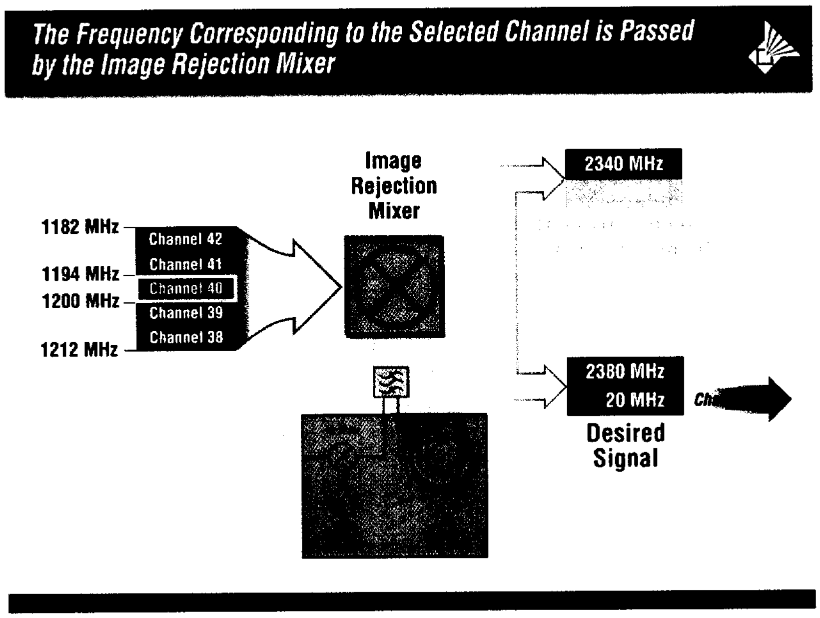
54. Since, as stated above, the television tuner has the entire television RF spectrum at its input, the output of the first mixer will contain a broad spectrum of signals at new frequencies. Because signals at frequencies other than the desired channel (channel 40 in this case) will cause interference and distortion in other parts of the tuner, it is advantageous to remove as many of these unwanted signals as possible. One method of accomplishing this is to use one or more filters between the mixers, typically of the bandpass variety. As depicted in Graphic Exhibit 19 below, the bandpass filter between the first and second mixers passes channel 40, the desired channel, as well as two channels above and below channel 40. All other channels are sufficiently attenuated, thus preventing interference in other parts of the tuner.

Graphic Exhibit #19



55. The output of the bandpass filter, consisting of five channels in this example, is the input to the second mixer, an image rejection mixer, in addition to the second local oscillator reference signal at a frequency of 1180 MHz, as depicted in Graphic Exhibit 20 below. Note that not all of the claims of the '035 patent require the second mixer to be an image rejection mixer. Unlike the variable-frequency first local oscillator, the second local oscillator is fixed at 1180 MHz. The second mixer will create two new signals from the interaction of the desired channel at 1200 MHz and the local oscillator at 1180 MHz. These two new signals are the difference component at 20 MHz (the desired second IF) and the sum component at 2380 MHz, which is not utilized.

Graphic Exhibit #20



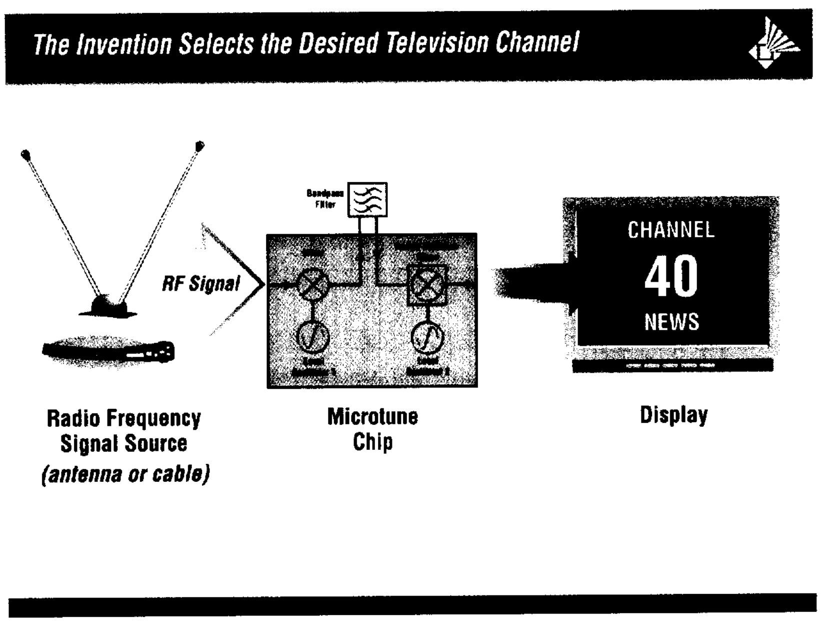
56. As shown in this example, the second mixer is an image rejection mixer, which is required for some but not all of the claims. The purpose of the image rejection mixer is to attenuate the image signal at 1160 MHz, which would be harmful because the 1160 MHz signal will combine with the local oscillator at 1180 MHz to produce a mixer output signal that is also at 20 MHz. Note that although this 1160 MHz image falls outside the five-channel passband of the bandpass filter discussed in the example above, in practice some signal energy remains at the image frequency, which is further attenuated by the image rejection mixer. This second 20 MHz signal would cause an undesired signal to be placed on top of the desired signal at 20 MHz and thereby cause interference with the desired signal. However, the image rejection mixer attenuates the image to reduce the interference. If the image were not properly attenuated, it would result in signal degradation, which, depending on the nature of the signal (television program or data), would cause visible picture disruption or data loss.

57. The degree of image rejection achieved by the image rejection mixer is complemented by the image rejection achieved by the bandpass filter. Note that the image

rejection mixer still produces the sum component associated with an input signal present at 1160 MHz, or 2340 MHz, but this is of little concern and easily filtered out of the mixer output.

58. Graphic Exhibit 21 below shows a high-level view of the operation of the television tuner. The large spectrum of television RF signals is selectively tuned until the output consists of only the desired channel, which is then further processed and decoded in subsequent circuitry to extract video and audio baseband signals in the case of analog television broadcasts, or to demodulate data in the case of digital television or cable modems.

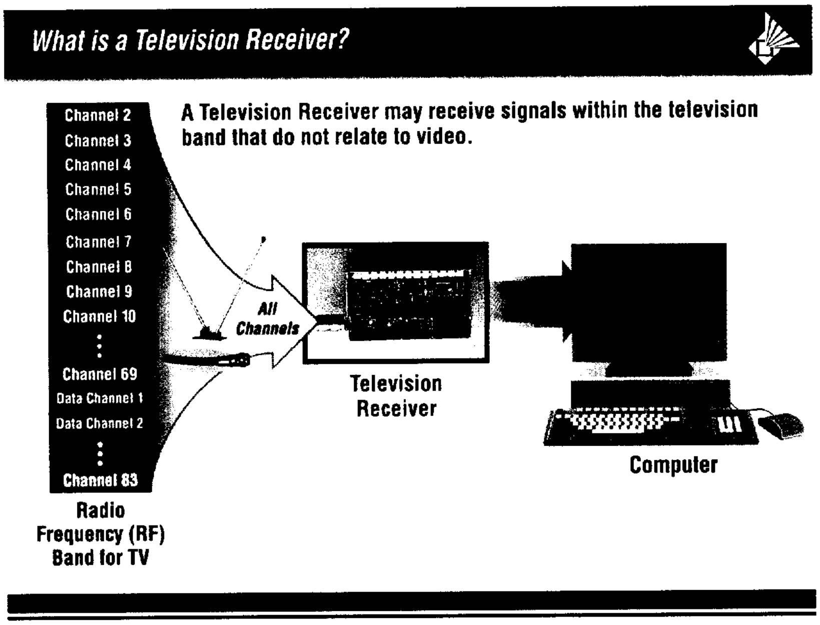
Graphic Exhibit #21



59. Although the operation of the television tuner inventions of the '035 patent have thus far been discussed in the context of a television set receiving video and audio signals, it is important to note that the television tuner disclosed in the '035 patent is advantageously used for other applications. For example, Graphic Exhibit 22 below shows the television tuner providing a tuned television signal to a computer for further processing. The television signal in this case would likely be comprised of digital data modulated in a Quadrature Amplitude Modulation (QAM) scheme. Standard 6-MHz television channels have been used to carry data in addition to

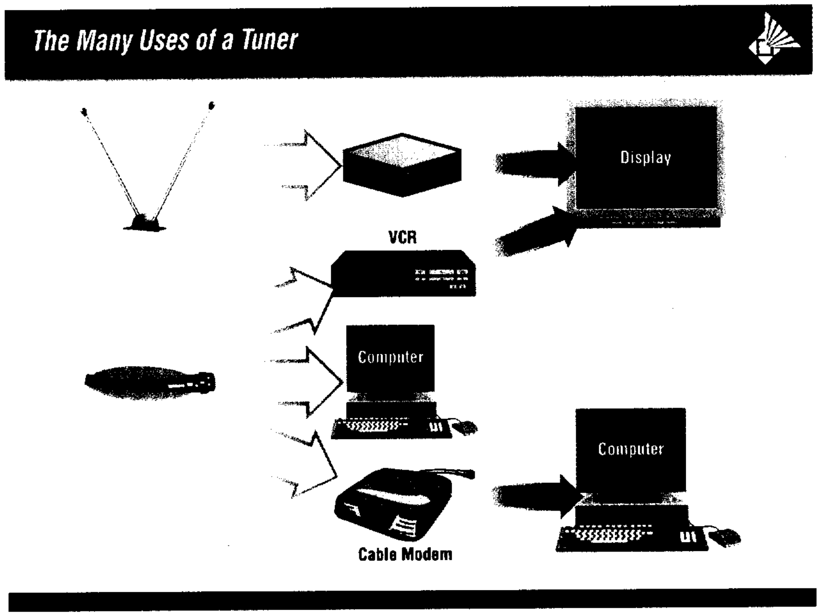
broadcast video/audio signals since the 1970s, and it is becoming commonplace to utilize television channels to carry data exclusively, such as digital cable telephony and two-way Internet connectivity for homes and businesses.

Graphic Exhibit #22



60. Various other uses for television tuners are shown in Graphic Exhibit 23 below. In addition to the television set and direct computer connections discussed above, the '035 television tuner would also be advantageously used in video cassette recorders (VCRs) and in cable modems, which convert QAM television signals into packetized digital data for use on computer networks. The scarcity of RF bandwidth available for new purposes assures that additional uses for conventional television signals will be found in the future. What all of these different uses of television signals share in common is the need for highly integrated, high performance, low-cost television tuners as disclosed in the '035 patent.

Graphic Exhibit #23



VI. CONSTRUCTION OF CLAIM TERMS.

61. I have reviewed the '035 patent and its file history to determine the proper claim construction for certain claim terms. It is my understanding that the claims are to be construed against the level of ordinary skill in the art, at the time of the invention, using the language of the claim, the specification, and, if necessary, the file history.

62. For purposes of claim construction, a person of ordinary skill in the art would have at least the following qualifications: a B.S. degree in electrical engineering, or the equivalent, and at least 2 years experience in the field of RF communications and/or television receiver circuit design.

63. **Television receiver.**

Definition.

The term "television receiver" means a device that receives and selects radio frequency (RF) signals within the television band, and is synonymous with "television tuner."

Support for Definition.

This term is used in the preamble of most of the '035 patent claims, and in the context of the '035 patent, this term is a device or circuit that receives the RF television signal, selects the desired channel, and tunes it such that it is converted to a standard intermediate frequency (IF), which is also known as a "television tuner." What the layperson calls a "television" or "television set" is really a combination of many different sub-components, for example, a television receiver to receive and tune the RF television signal, circuits that process the decoded video and audio signals, a high-voltage power supply, and a picture tube to actually display the received video signal. In this day and age in which television channels are not just carrying broadcast stations, but also data and connectivity to the Internet, it is important to remember that a television receiver is not restricted to use in a television set, but can also be found within a personal computer (PC) or network interface device such as a cable modem. Regardless of the intended use of the television channel itself, the function of a television receiver in each of these examples is the same.

I have reviewed the prosecution history. The U.S. Patent and Trademark Office examiner referred to the invention of U.S. Patent No. 4,162,452 to Ash (the "Ash reference") (Exhibit 11) as a "television receiver." Exhibit 3, at 5. Because the Ash reference deals with tuning television stations, the reference by the U.S. Patent and Trademark Office ("PTO") to Ash as a "receiver" supports and is consistent with my definition.

64. **Image, Image Signal.**

Definition.

The terms "image" and "image signal" are synonymous and mean an undesired signal that, along with the desired signal, is present at the signal input of a mixer. The image signal is located at a frequency that is twice the frequency of the desired mixer output signal, namely the intermediate frequency (IF), away from the frequency of the desired signal. The result, in the

case of a basic mixer, is to create a second output signal at the same frequency as the desired output signal, thereby causing interference with the desired mixer output signal.

Support for Definition.

The frequency of an image signal for a given desired mixer input signal frequency and the corresponding local oscillator input signal frequency can be determined by using a formula derived from the definition above. When the desired mixer output signal is the difference component, if the local oscillator frequency is higher than the frequency of the desired signal, the frequency of the image signal will be higher than the frequency of the desired signal; if the local oscillator frequency is lower than the frequency of the desired signal, the frequency of the image signal will be lower than the frequency of the desired signal. The formula is: f_{IM}=(2 x f_{LO})- f_{RF}, where f_{IM} is the image signal frequency, f_{LO} is the local oscillator frequency and f_{RF} is the desired input signal frequency. The specification provides ample support for these terms. For example, it is stated that "[t]he image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF}=|1160-1180|=20 MHz." Ex. 1, at col. 10, ll. 4-7. Note that application of the formula f_{IM}=(2 x f_{LO})- f_{RF} yields f_{IM} =(2 x 1180) – 1200 = 1160 MHz, the image signal frequency identified in the specification. The term "image" is used synonymously and consistently with "image signal" in the specification, as in "the image falling on 1160 MHz." Col. 10, l. 30. Similarly, during prosecution the applicant used the term "image signal" in reference to the 1160 MHz image to the examiner that "MIX2 [the image rejection mixer] must reject these image signals associated with the first IF signal." Ex. 5, at 9. Thus, the terms "image" and "image signal" are properly defined in the specification and used consistently throughout the specification and claims.

65. **Image Rejection, Rejects an Image Signal.**

Definition.

The term "image rejection" and the phrase "rejects an image signal" refer to attenuation of an image signal to such a degree as to minimize its undesired interaction, and thereby reduce interference with the desired output signal.

Support for Definition.

Although the '035 patent specification does not explicitly define the term "image rejection" or the phrase "rejects an image signal," I find that the concept of "image rejection," namely that an image signal is rejected or significantly attenuated to prevent interference with a desired signal, is adequately described in the specification, and that the definition is consistent with image rejection, as it is known in the art. There are two common methods for performing image rejection, which can be used individually or together for cumulative effect. The first is by filtering, typically with a bandpass filter, such that the image signal falls somewhere outside the passband of the filter. The second method is to use an image rejection mixer, which is a mixer that is specifically designed to pass the desired frequency while attenuating the image signal. When both an image rejection mixer and a filter are used, the total image rejection achieved (usually expressed in decibels, or dB) is the sum of the image rejection performed by each.

The '035 patent discloses three methods of performing image rejection. Figure 4 shows using an image rejection mixer alone. The description of Figure 4 discloses that a filter could be used in conjunction with the image rejection mixer. *See id.* at col. 12, ll. 44-46 ("The design will include a bond wire parallel LC network option to further band-limit any image energy at the output of this mixer if necessary.") In context, this discussion suggests to one of ordinary skill in the art that a "parallel LC network" is placed at the output of the first mixer to prevent energy at the image frequency from reaching the input of the second mixer. A "parallel LC network" consists of an inductor and capacitor in parallel and forms a bandpass filter, as known in the art. A "bond wire" is a wire attached to an IC that acts as the inductor in the bandpass filter.

As with the "bond wire" image rejection filter, Figure 5 shows a filter 509 between two mixers. The symbol for the second mixer is general and does not specify an image rejection type of mixer. However, for improved image rejection as discussed with respect to Figure 4 (above) the second mixer in Figure 5 could be an image rejection mixer.

Exhibits 8-10 relate to bond wire filters. These exhibits are technical articles and confirm that a bond wire has suitable characteristics to be used as the inductor in a bandpass filter in the range of frequencies relevant to the '035 invention. These articles indicate that bond wire filters possess suitable values to perform image rejection, if desired.

66. **Image Rejection Mixer, Image-Rejection Mixer.**

Definition.

The terms "image rejection mixer" and "image-rejection mixer" are synonymous and refer to a certain type of mixer that performs image rejection in addition to combining input signals. In other words, an image rejection mixer not only performs the functions of a basic mixer, but also specifically attenuates the image signal to limit interference with the desired mixer output signal.

Support for Definition.

In my opinion, the '035 patent specification as originally filed contains both sufficient definition of the necessity for and function of an image rejection mixer, and further provides an example of how to construct an image rejection mixer that would function appropriately to perform image rejection. The specification first defines an image rejection mixer as having characteristics similar to a basic mixer, namely that "[t]he IF frequencies, given by $f_{IF} = |f_{RF} \pm f_{LO}|$, have carriers at f_{IF}=1200-1180=20 MHZ and at f_{IF}=1200+1180=2380 MHZ." *Id.* at col. 10, ll. 1-3. The specification then states that "[t]he image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF} =|1160-1180|=20 MHz." *Id.* at col. 10, ll. 4-7. One example of an image rejection mixer is provided in Figure 6, and its operation described in the specification at column 13, lines

1-14. The image rejection mixer disclosed in Figure 6 is just one method of constructing an image rejection mixer, and from my reading of the '035 patent claims, there are many other image rejection mixer designs that would necessarily fall within the definition of image rejection mixer as used in the '035 patent.

67. **Subtractive Mixer, Subtractive Mixing, Subtractively Mixing, Subtractively Mixes.**

Definition.

A "subtractive mixer" is a mixer in which only the frequency difference signal components created by the combination of the two mixer input signals are present in the output signal, the frequency sum components being removed by a filter. "Subtractive mixing," "subtractively mixing" and "subtractively mixes" means combining two mixer input signals in a mixer such that only the frequency difference signal components created by the mixer are present in the output signal, the frequency sum components being removed by a filter.

Support for Definition.

"Subtractive mixer" is defined in the specification by noting how it differs from a basic mixer. The specification illustrates the operation of a subtractive mixer by first pointing out that a basic mixer with a 400 MHz signal at its RF input and a local oscillator input of 1600 MHz will create two mixer outputs, one at 2000 MHz, and one at 1200 MHz. *See id.* at col. 9, ll. 23-28. It is then identified that "[s]ince mixer 408 is a subtractive mixers [sic] the 2000 MHz signal will be filtered out and only the 1200 MHz signal will be passed on to mixer 410," illustrating the difference between the "subtractive mixer" and basic mixer. *Id.* at col. 9, ll. 30-32.

The phrases "subtractive mixing," "subtractively mixing" and "subtractively mixes" should be construed consistently with the operation of combining two inputs in a "subtractive mixer," as defined above.

68. **Filter, Filters, Filtered, Filtering.**

Definition.

A "filter" is defined as a device that passes signals of certain frequencies and attenuates signals of other frequencies. "Filters," "filtered" and "filtering" means processing a signal such that signals of certain frequencies are passed and signals of other frequencies are attenuated.

Support for Definition.

The term "filter," as used in the claims and specification of the '035 patent, is completely consistent with the concept and definition of a filter as used in the art. The specification describes several filters used in both the prior art television tuner and the integrated television tuner of the '035 patent, and in each case, the recited filters pass certain frequencies while attenuating others. *See id.* at col. 5, ll. 50-52, col. 5, ll. 54-59. Moreover, the verb forms "filters," "filtered" and "filtering" as used in various claims are consistent with the definition of "filter" and are therefore fully supported by the specification.

69. **Bandpass Filter, Passband.**

Definition.

A "bandpass filter" is a filter that passes signals that lie within a selected band of frequencies with low loss, called the "passband," but attenuates signals at frequencies above and below the "passband."

Support for Definition.

The specification describes the operation of a representative bandpass filter as "[f]ilter 304 also attenuates signals not in a fairly narrow (100 MHz) range about the desired signal." *Id.* at col. 5, ll. 54-55. In this example, the "passband" is 100 MHz wide, and the bandpass filter attenuates frequencies both below and above this passband, as indicated by the phrase "about the desired signal." The specification also refers to upper and lower "passband" limits in describing the upper and lower frequency limits of the passband of Second IF filter 420, which is of the

bandpass type. *See id.* at col. 13, ll. 51-53. The usage of both bandpass filter and passband are consistent in their usage throughout the claims.

70. **Receiver input "coupled to an RF signal source," Input filter "coupled to an RF signal source."**

Definition.

The phrase "coupled to an RF signal source" means that an electrical connection is made to a source of one or more RF signals, for example, an antenna receiving over-the-air signals or a coaxial cable for connection to a cable television system. In my view, the term "RF signal source" is not a limitation, and an RF signal source need not be included in an accused product, but an accused product must, in operation, be "coupled" to an RF signal source.

Support for Definition.

Although this phrase is not defined in the specification, it is stated that the prior art "[t]elevision tuner 300 receives a standard television RF signal from either antenna 302 or a cable system connection (not shown)." *Id.* at col. 5, ll. 47-49. In the '035 invention, "[t]he RF signal enters television tuner 100 from the antenna 402 (or cable, not shown) and is passed through RF low-pass filter (RFLPF) 404." *Id.* at col. 8, ll. 32-34. Thus, the specification provides examples of two ways of coupling the television tuner to an RF signal source: an antenna or connection to a cable television system. I believe that these RF signal sources were only meant to be examples, and that any source of RF signals should be construed as possible within the claims.

71. **Passing more than one channel.**

Definition.

The phrase "passing more than one channel" means that the passband of the bandpass filter processing a television RF signal would be greater than one channel, i.e., greater than the 6-MHz bandwidth of a standard television channel in the United States.

Support for Definition.

It is well known in the art, and the specification states, "each channel in the VHF and UHF band has a 6-MHz bandwidth." *Id.* at col. 8, ll. 43-44. "Passing more than one channel" must necessarily mean passing more than 6 MHz. I note that several prior art references cited by the Examiner during prosecution of the '035 patent, including the Ash reference (Exhibit 11) and U.S. Patent No. 4,340,975 to Onishi *et al.* (the "Onishi reference") (Exhibit 12), employed bandpass filters with a passband of one channel or less. The Onishi reference discloses a passband of 5 MHz, and the Ash reference discloses a passband of 6 MHz. *See* Ex. 11, at col. 5, ll. 65-68; Ex. 12, at col. 10, ll. 55-59.

72. **Having at least four channels.**

Definition.

The phrase "having at least four channels" means that the passband of a bandpass filter processing a television RF signal would be greater than four television channels, i.e., 24 MHz, the bandwidth of four contiguous standard television channels in the United States.

Support for Definition.

The phrase "Having at least four channels" must necessarily mean passing more than 4 television channels. For example, the specification states, "each channel in the VHF and UHF band has a 6-MHz bandwidth." Ex. 1, at col. 8, ll. 43-44. Four times 6 MHz is 24 MHz. I note that several prior art references cited by the Examiner during prosecution of the '035 patent, including the Ash and Onishi references, employed bandpass filters with a passband of one channel or less. The Onishi reference discloses a passband of 5 MHz, and the Ash reference discloses a passband of 6 MHz. *See* Ex. 11, at col. 5, ll. 65-68; Ex. 12, at col. 10, ll. 55-59.

73. **A first bandpass filter...performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent circuitry.**

Definition.

The phrase "a first bandpass filter...performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent circuitry" means that the bandpass filter accomplishes some detectable measure of image rejection, but less than all of the image rejection accomplished by the invention, and further that the bandpass filter decreases the total power content of the signal it processes, thus reducing the power-handling requirements of circuitry that follows the bandpass filter.

Support for Definition.

Filters commonly used between the first and second mixers of television tuners for the purpose of rejecting the image frequency at the input to the second mixer are typically designed to have a narrow passband, such as one channel or 6 MHz. A narrow passband filter usually has greater attenuation within its passband than does a wider passband filter. The latter effect results in greater loss of energy of the desired signal. In reviewing several prior art patents cited by the Examiner during prosecution of the '035 patent, I noted that the prior art television tuners disclosed in the Ash and Onishi references each had bandpass filters that accomplished all of the image rejection, meaning no other component was responsible for image rejection, and that the passbands of the respective filters were quite narrow (one channel or less), as expected. *See* Ex. 11, at col. 7, l. 67 – col. 8, l. 4; Ex. 12, at col. 10, ll. 55-59. In contrast to the bandpass filters disclosed in these prior art television tuners, the bandpass filter recited in claim 83 of the '035 patent performs *some* of the image rejection; with the balance performed by the second mixer, which is an image rejection mixer.

74. **"Coupled to" (claim 120).**

Definition.

The phrase "a second mixer having a first input coupled to an output of said first mixer" means that the signal path extends from the first mixer output to the second mixer output, and that there can be intermediate circuit components between the first mixer output and second mixer input.

Support for Definition.

The plain meaning of "coupled to" means "connected to." However, an electrical component can be connected to another electrical component even if there are intermediate components connected in between. In fact, the specification shows a first mixer 408 connected to an intermediate component (a filter 509) and then, connected to a second mixer 510. *See* Figure 5. The specification, in Figure 4, also shows a "direct" connection between mixers 408 and 410 without a filter between the mixers.

The "coupled to" language was present in originally filed claim 1. In its first office action, the PTO construed and rejected claim 1. Claim 1 was rejected over the Ash reference, and construed to permit a component (a filter in the case of Ash) to exist between the first mixer output and the second mixer input. *See* Ex. 3, at 2 (rejection). In response, the applicant amended claim 1 to recite "coupled directly to." *See* Ex. 4, at 2 (insertion of "directly" into claim 1); *id.* at 7-8 (discussion of addition of word "directly"). This amendment was made to overcome the PTO's construction of claim 1 as permitting intermediate components.

When adding claim 121, now claim 120, the applicant did not use "coupled directly to" but instead kept the original broader wording of "coupled to." Claim 120 was filed during the prosecution of the patent (as claim 121 at that time) after the examiner had already construed "coupled to" broadly. *See* Ex. 7, at 2 (adding claim 121, now renumbered as claim 120). The applicant argued that claim 120 was patentable due to the limitations present in allowed claims

10 and 28, which were also inserted into claim 120. Those limitations concerned an "input filter" and an image rejection mixer. *See id.* at 4:

> New independent claim 121 [now claim 120] includes both the limitations of an input filter and an image rejection mixer. Applicant submits that the prior art does not teach or suggest the combination of elements in claim 121, and therefore, claim 121 is patently novel and nonobvious.

Based on the claim language, specification and file history, the claim phrase "coupled to" in claim 120 should be construed broadly to permit the inclusion of other circuit components between the two mixers, such as a filter.

75. I prepared this declaration to aid the Court in construing the claim terms of the '035 patent. In doing so, I endeavored to include only as much detail as I thought would be helpful, while omitting some information that, although technically accurate, might make it more difficult to understand the important basic concepts. I may amend, change or provide further detail if necessary to aid the Court in the construction of the claim terms.

I declare under penalty of perjury under the laws of the United States and the State of Texas that the foregoing is true and correct and that this declaration was executed on January 20, 2002, at Princeton, New Jersey.

Bernard J. Lechner

BERNARD J. LECHNER

CERTIFICATE OF SERVICE

I certify that a true and correct copy of the above Declaration was sent to the following, on January 22, 2002, via:

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TAB 1

US005737035A

United States Patent [19]

Rotzoll

[11] Patent Number: 5,737,035

[45] Date of Patent: Apr. 7, 1998

[54] HIGHLY INTEGRATED TELEVISION TUNER ON A SINGLE MICROCIRCUIT

[75] Inventor: Robert Rudolf Rotzoll, Allen, Tex.

[73] Assignee: Microtune, Inc., Plano, Tex.

[21] Appl. No.: 426,080

[22] Filed: Apr. 21, 1995

[51] Int. Cl.[6] H04N 5/44

[52] U.S. Cl. 348/725; 348/731; 455/315; 455/339

[58] Field of Search 455/302, 315, 455/339, 179.1; 348/725, 726, 731; H04N 5/44

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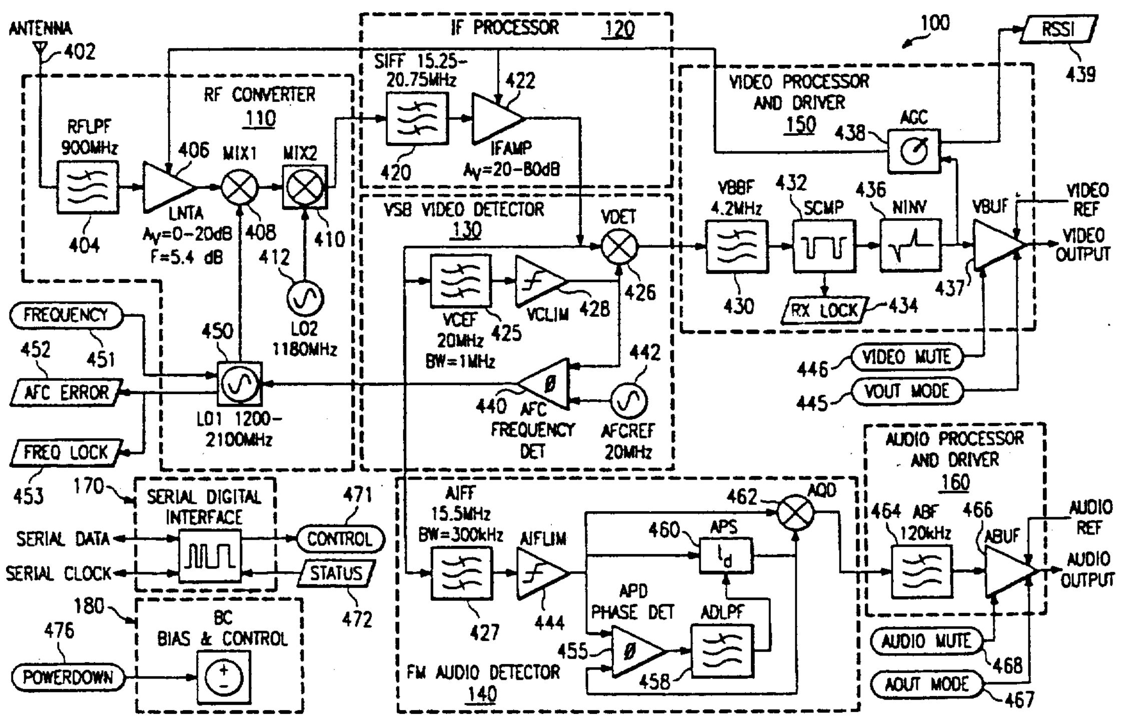
Primary Examiner—Sherrie Hsia
Attorney, Agent, or Firm—Fulbright & Jaworski L.L.P.

[57] ABSTRACT

There is disclosed a fully integrated television receiver for receiving a standard antenna or cable input and outputting a standard video baseband signal and a standard audio baseband signal. The receiver employs an up-conversion mixer and a down-conversion image-rejection mixer in series to produce an on-chip IF signal. Audio detection of the IF signal is performed via an audio delay-locked loop operating in quadrature to the audio FM signal. Video detection of the IF signal is performed by frequency mixing the IF signal with the extracted video carrier of the IF signal. Variable load drive capability is provided for both the video output and the audio output.

128 Claims, 4 Drawing Sheets



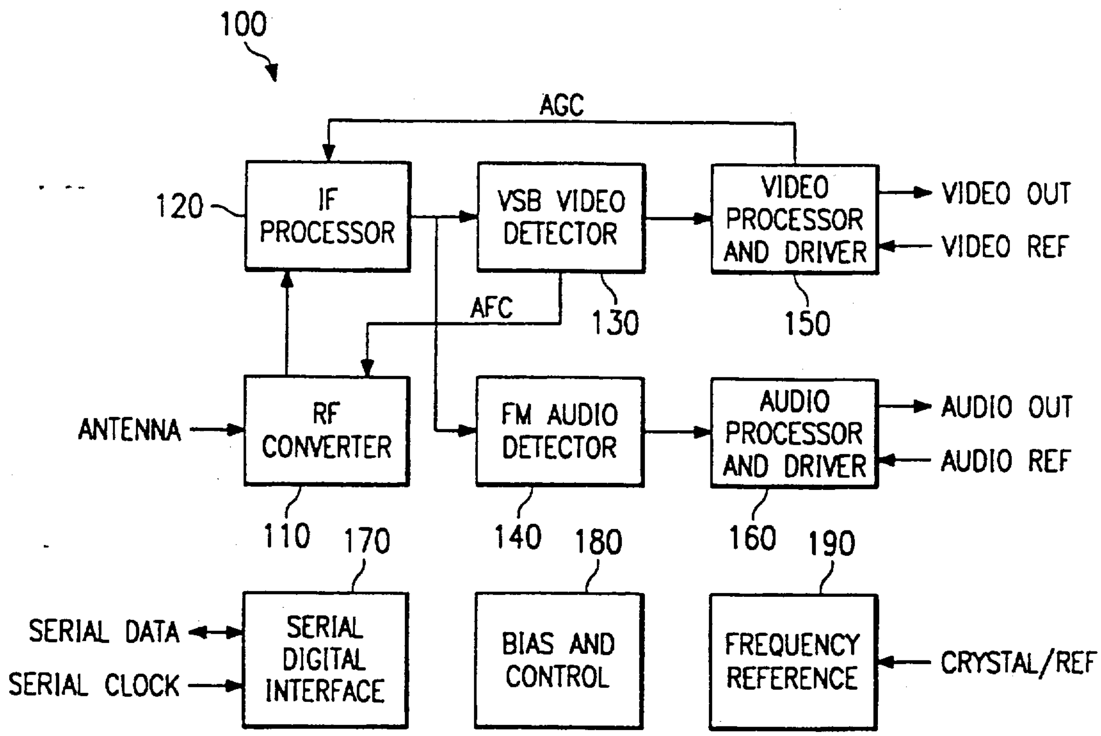

100
AGC
120
IF PROCESSOR
VSB VIDEO DETECTOR
VIDEO PROCESSOR AND DRIVER
VIDEO OUT
VIDEO REF
AFC
130
150
ANTENNA
RF CONVERTER
FM AUDIO DETECTOR
AUDIO PROCESSOR AND DRIVER
AUDIO OUT
AUDIO REF
110
170
140
180
160
190
SERIAL DATA
SERIAL CLOCK
SERIAL DIGITAL INTERFACE
BIAS AND CONTROL
FREQUENCY REFERENCE
CRYSTAL/REF

FIG. 1

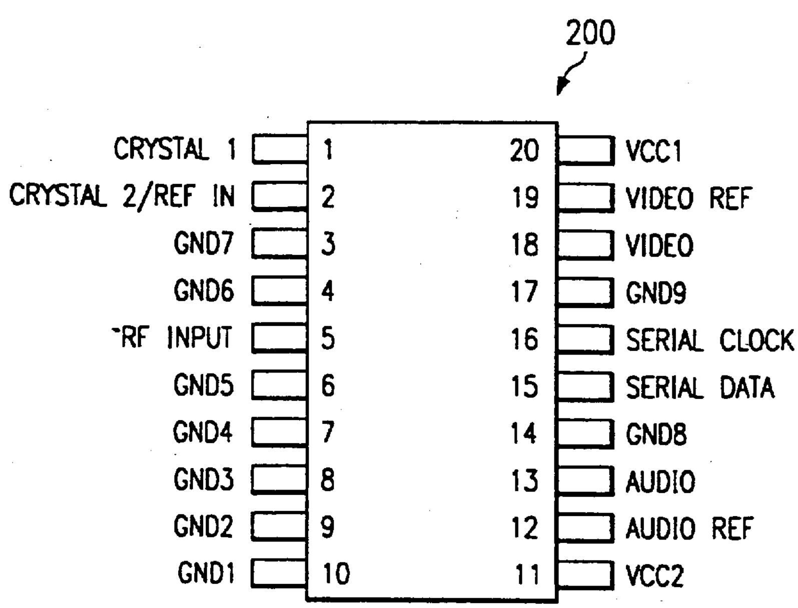

200
CRYSTAL 1
CRYSTAL 2/REF IN
GND7
GND6
RF INPUT
GND5
GND4
GND3
GND2
GND1
1
2
3
4
5
6
7
8
9
10
20
19
18
17
16
15
14
13
12
11
VCC1
VIDEO REF
VIDEO
GND9
SERIAL CLOCK
SERIAL DATA
GND8
AUDIO
AUDIO REF
VCC2

FIG. 2

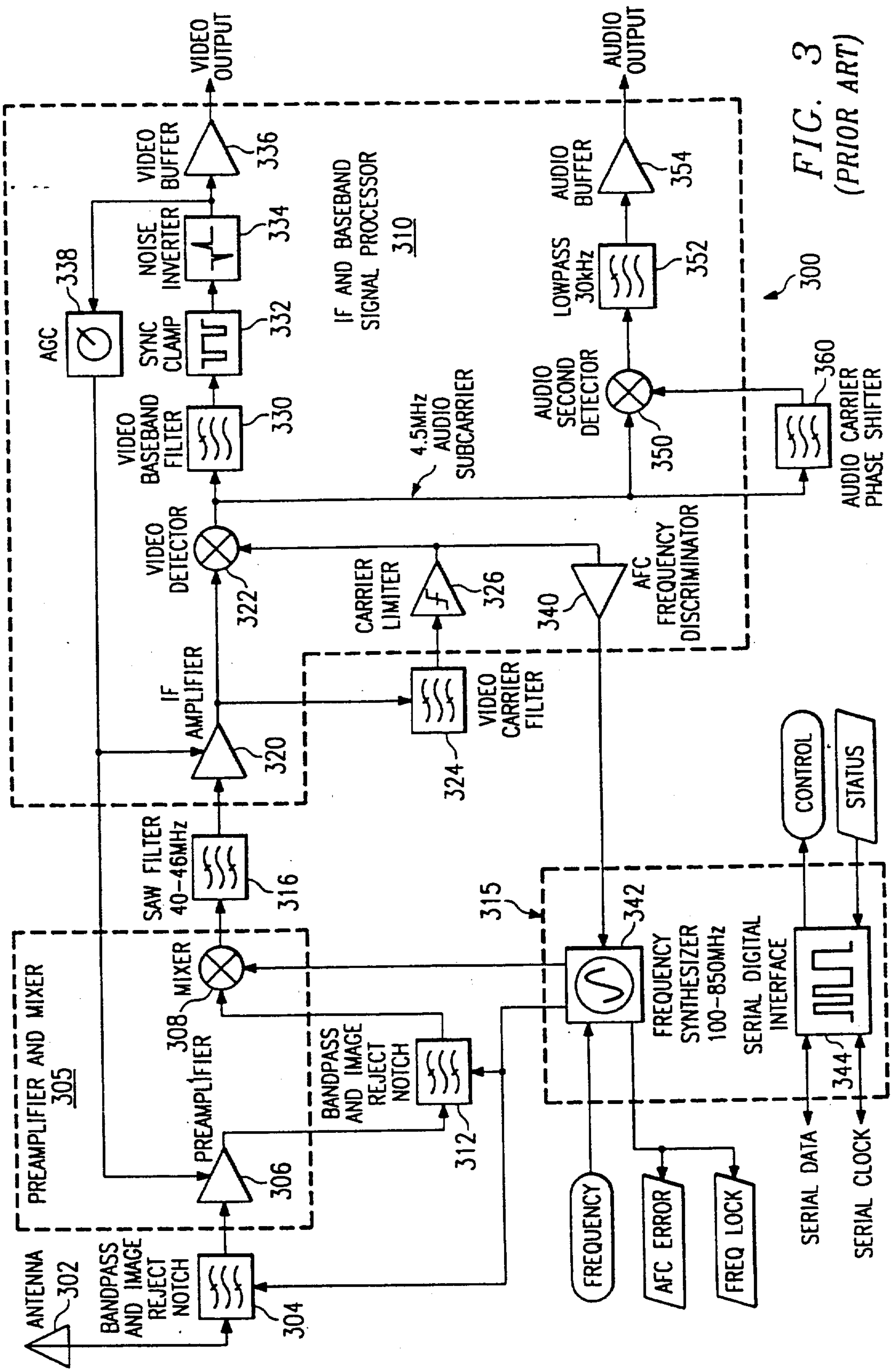


FIG. 3 (PRIOR ART)

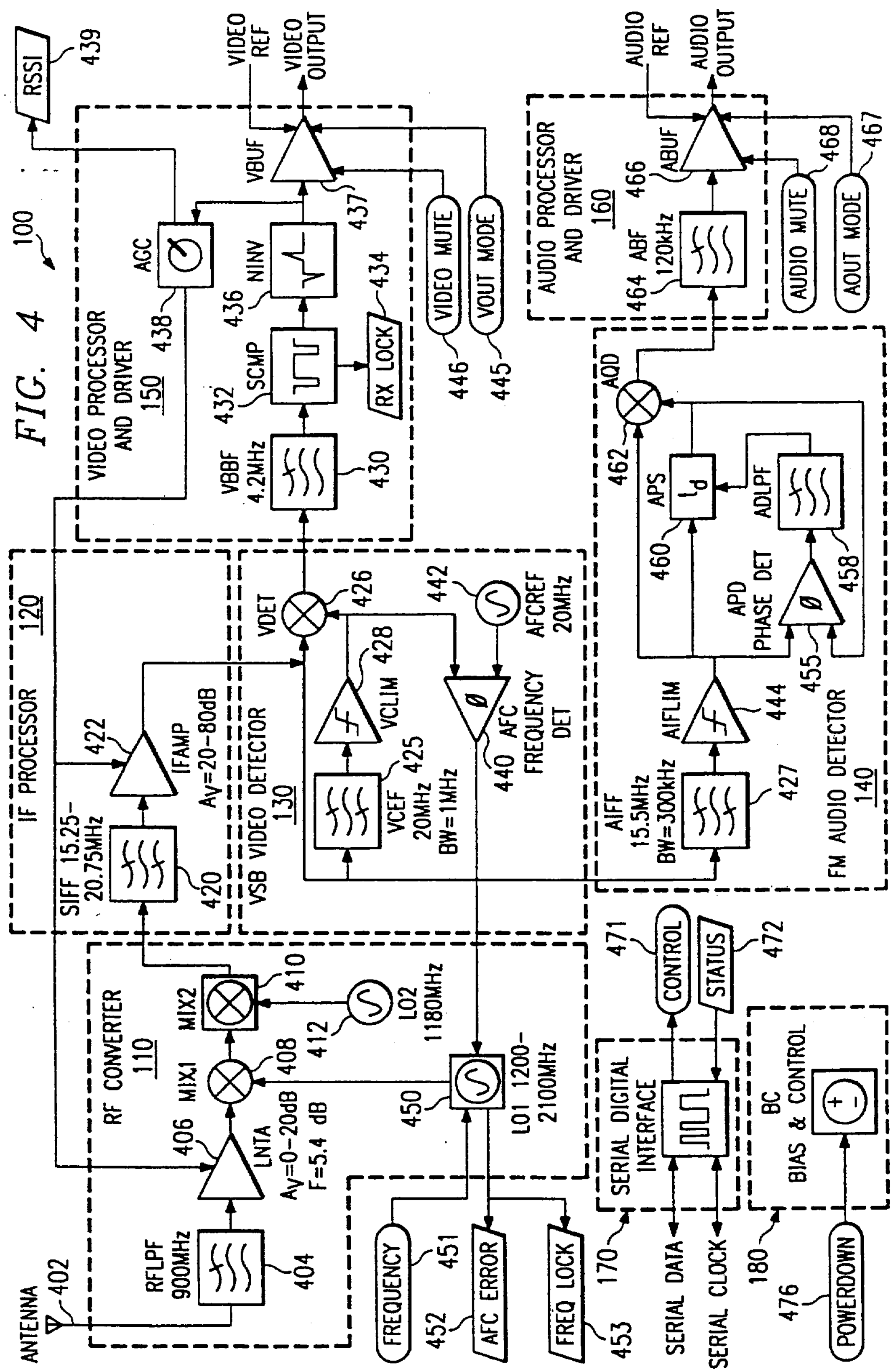


FIG. 4

TUNE 036947

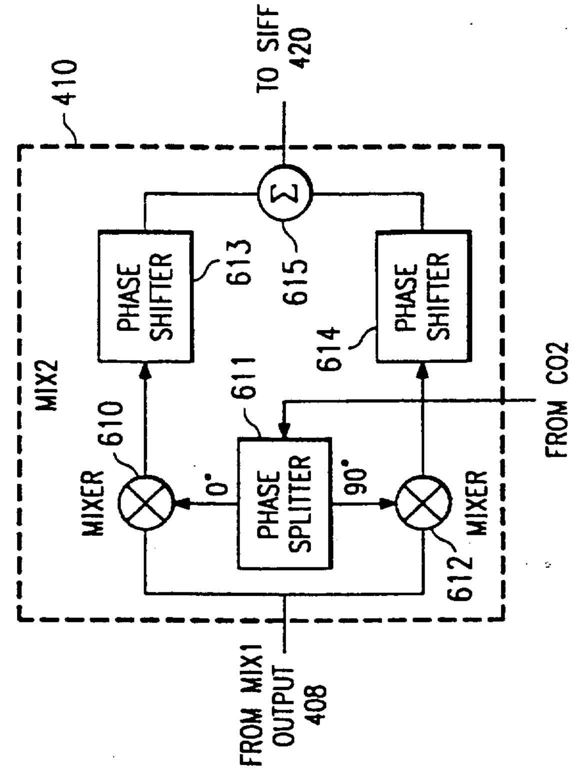

410
TO SIFF
420
MIX2
PHASE SHIFTER
613
615
Σ
PHASE SHIFTER
614
MIXER
610
611
0°
PHASE SPLITTER
90°
MIXER
612
FROM CO2
412
FROM MIX1 OUTPUT
408

FIG. 6

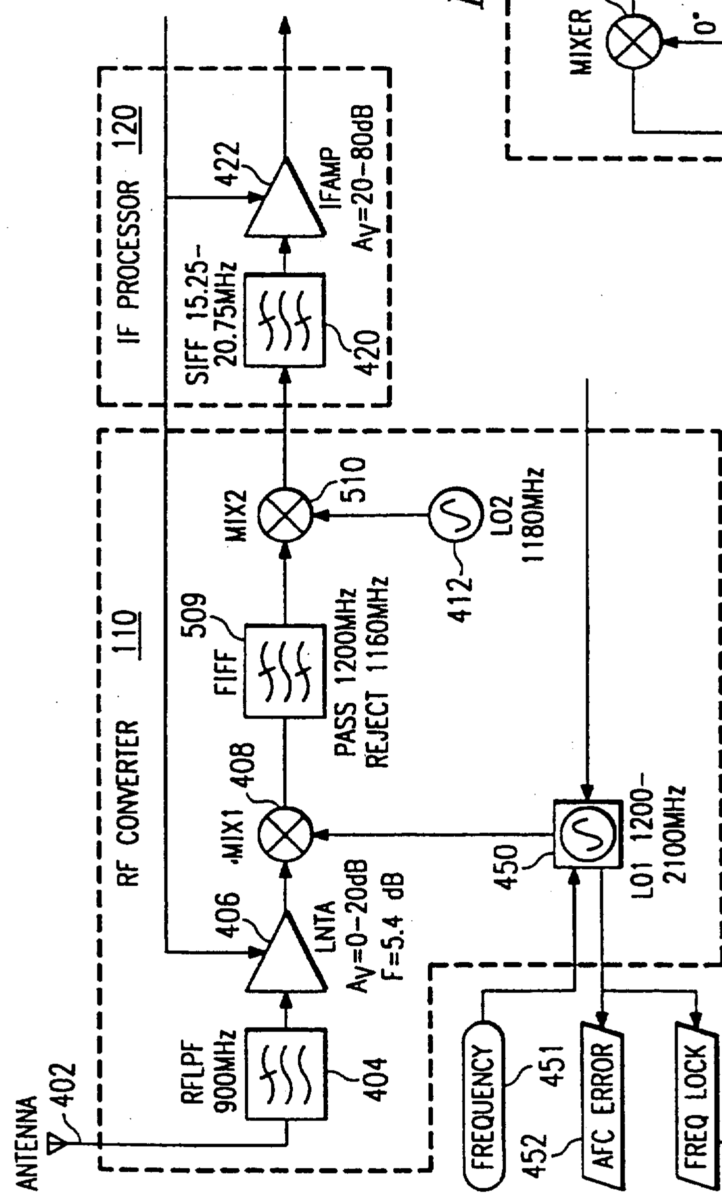

IF PROCESSOR 120
SIFF 15.25–20.75MHz
420
422
IFAMP
Av=20–80dB
RF CONVERTER 110
ANTENNA
402
RFLPF 900MHz
404
406
LNTA
Av=0–20dB
F=5.4 dB
MIX1
408
509
FIFF
PASS 1200MHz
REJECT 1160MHz
MIX2
510
412
LO2
1180MHz
450
LO1 1200–2100MHz
FREQUENCY
451
452
AFC ERROR
FREQ LOCK
453

FIG. 5

HIGHLY INTEGRATED TELEVISION TUNER ON A SINGLE MICROCIRCUIT

TECHNICAL FIELD OF THE INVENTION

This invention relates to television tuner circuits and more particularly to a highly integrated television tuner fabricated in a single microcircuit device.

BACKGROUND OF THE INVENTION

One of the most significant costs in television manufacturing is the cost of the tuner. The typical cost of a television (TV) tuner is in the neighborhood of $15.00. which. relative to the cost of the entire television set. is very substantial. Part of the solution to reducing tuner cost is to reduce the number of components in the tuner.

Traditionally. tuners have been comprised of two basic components. The first component performs high frequency to intermediate frequency (RF to IF) conversion. Subsequently. the second component performs IF to baseband conversion. The TV tuner was originally designed for broadcast television reception within a television set. which is essentially a stand-alone unit containing a cathode ray picture tube. So. TV tuners were originally integral parts embedded in a single-purpose device.

Presently. however. state-of-the-art consumer electronic devices use TV tuners that are not a built-in part of a television set. The tuner is a separate element that is connected to a cathode ray picture tube at some point. but the tuner is not an integral part of the monitor. For example. TV tuners may be fabricated on circuit boards and then installed in personal computer (PC) systems. thereby allowing the PC to function as a television set. These tuners convert a radio frequency television signal into a baseband (or low frequency) video signal which can then be passed on to other elements in the PC for video processing applications.

The circuit component that performs the RF-to-IF conversion typically comprises one or two integrated circuits and numerous discrete elements—inductors, capacitors and/or transistors. The IF-to-baseband conversion typically includes another integrated circuit. several filter elements. such as ceramic filters and SAW filters. a series of tuning and control elements. such as resistors and potentiometers. variable inductors and/or capacitors. and some other additional external components. Thus. the complexity of the tuner is fairly high and typically there may be between 100 and 200 elements on a circuit board. Furthermore. state-of-the-art TV tuners still require that each tuner be aligned by manual tuning before leaving the factory. This manual tuning is one of the most expensive costs associated with the manufacturing process and an important factor in the cost of tuners.

Broadcast television tuners of the past have gone through an evolution over a period of more than 60 years. The earliest tuners utilized vacuum tube technology and required that the minimum number of vacuum tubes possible be used due to their cost. power consumption and dimensions. Therefore. passive components. such as resistors. capacitors. inductors and transformers. were used as much as possible in most designs. This style of design continued until about 1960 when TV tuner components. particularly vacuum tubess began to be replaced by bipolar and MOS transistors. However; the active device count still defined the cost and size limits of TV tuners and active device count minimization continued.

In the early 1970's the integrated circuit became viable as an element in the television tuner and the design techniques

were dramatically changed. Many functions of the tuner utilizing only one tube or transistor were being replaced with 4 to 20 individual transistors which could perform the same function with better precision. less spaced less powers less heat generation and lower costa The introduction of the integrated circuit was gradual. first encompassing only low frequency elements and then eventually high frequency active elements. Nonetheless. many passive elements external to the integrated circuits remained in TV tuner designs.

One advance. the SAW (surface acoustic wave) filters made a significant change in that several manually tuned inductors and capacitors could be removed from the tuners and receive-filtering performance could be improved within a much smaller space and at reduced cost. However. the SAW filter. which is fabricated on a ceramic substrate. cannot be integrated on a silicon wafer with the rest of the active circuitry and must therefore remain a discrete component in the final design. The trend of the 1980's was to miniaturize all of the passive components and simplify their associated manual tuning at the factory. In recent years. TV tuners have been reduced in size from requiring fairly large enclosures. about 2"×5"×1". to much smaller enclosures. about ½"×2"×⅜". There is a high premium placed on small size because TV tuners are being used in smaller and smaller computers. television sets and VCRs. As the equipment in which tuners are used becomes smaller. the size of the TV tuner must decrease also.

As the size of the tuner goes down. and as tuners are used in a wider variety of devices. cost becomes more critical and must be reduced as much as possible in order not to represent a large portion of the final product cost. When a tuner is used in a television set. the tuner size is less critical because the television set inherently has a large mass. But when a tuner is used in other electronic equipments space becomes a premium and the footprint of the tuner becomes critical.

Accordingly. it is one object of the invention to provide a TV tuner which has a relatively low cost and a small footprint for use on a printed circuit board.

It is another object of the present invention to provide a TV tuner that meets or exceeds the performance of state-of-the-art TV tuners while at the same time reducing the number of external components needed. thereby decreasing the complexity of the printed circuit board and the amount of circuit board area needed by the TV tuner.

It is the further object of the present invention no allow for computer control of the TV tuner by a serial bus so that the TV tuner may be controlled by a microcontroller imbedded in the television set. personal computer. or other video device.

It is the further object of the present invention to provide a TV tuner with computer-controlled output impedance characteristics to accommodate different load specifications.

SUMMARY OF THE INVENTION

These and other problems have been solved by integrating a TV tuner that takes a broad band of frequencies as an input and performs video demodulation over a very narrow band width. To accomplish this. an architecture was chosen to perform an up-conversion of the RF input signal to a higher internal frequency. which allows the present invention to have minimal filtering on the input stages of the receiver. The present invention is therefore able to operate without variable tuned filtering on its input. with the exception of a single fixed tuned filter. This eliminates the need for precisely controlled variable tuned filters which must be

made to the following descriptions taken in conjunction with the accompanying drawings, in which:

FIG. 1 is a high level block diagram of the functional components of the present invention;

FIG. 2 is a pin layout diagram of an integrated circuit in accordance with the present invention;

FIG. 3 is a detailed block diagram of a state-of-the-art television tuner found in the prior art;

FIG. 4 is a detailed block diagram of the circuit components of the present invention;

FIG. 5 shows an alternate embodiment of the RF front-end of the present invention; and

FIG. 6 is a detailed block diagram of MIX2 of FIG. 4.

DETAILED DESCRIPTION OF THE PRIOR ART

Before discussing the monolithic television tuner of the present invention, it will be useful to discuss a state-of-the-art television tuner found in the prior art.

While there have been theoretical proposals to integrate TV tuners in a single microcircuit, none are known to have been implemented. The next best definition of the known prior art, then, is a highly miniaturized, but not fully integrated, tuner as shown in FIG. 3.

FIG. 3 depicts a functional electrical block diagram of a present state-of-the-art TV tuner configuration. Television tuner **300** is constructed in a single metallically shielded assembly containing a printed circuit board on which all of the associated tuner components are mounted and electrically connected. TV tuner **300** is designed to be a module mounted on other printed circuit boards to allow for direct connection of the input and output signals to their appropriate terminations within the television receiving system. The metal shield is used to keep undesired external signals from interfering with the operation of the TV tuner **300** and to prevent TV tuner **300** from radiating signals that interfere with the operation of external devices.

Prior art television tuner **300** is comprised of three integrated circuits: preamplifier and mixer **305**, IF and baseband signal processor **310** and frequency synthesizer and Inter integrated Circuit (IIC or I^2C) bus interface **315**. Television tuner **300** is also comprised of a plurality of discrete components, including bandpass and image reject notch filter **304**, bandpass and image reject notch filter **312**, surface acoustic wave (SAW) filter **316**, video carrier filter **324**, and audio carrier phase shifter **360**.

Television tuner **300** receives a standard television RF signal from either antenna **302** or a cable system connection (not shown) through bandpass and image reject notch filter **304**. Bandpass and image reject notch filter **304** limits the signals entering TV tuner **300** so that a minimum number of undesired signals exist in TV tuner **300**. Filter **304** therefore limits the image response caused by the first mixer, described later. Filter **304** also attenuates signals not in a fairly narrow (100 MHz) range about the desired signal. Finally, known interference signals, such as FM broadcast, shortwave service signals, signals in the intermediate frequency band and Citizen Band radio signals, are specifically rejected by filter **304**.

Preamplifier **306** of preamplifier and mixer **305** receives the output of bandpass and image reject notch filter **304** and raises the signal level (10 dB) with minimum increase in the noise level (typically 8–10 dB). The gain of preamplifier **306** is controlled by automatic gain control (AGC) **338**, so that when a very strong signal enters TV tuner **300**, overall gain is reduced, resulting in less distortion in the preamplifier than without the gain reduction.

The output of preamplifier **306** is sent to bandpass and image reject notch filter **312**, with the same basic requirement of minimizing the passage of potential interference signals. Filter **312** is external to preamplifier and mixer **305** and is comprised of a plurality of discrete elements, including capacitors, inductors and varactor diodes.

The output of bandpass and image reject notch filter **312** is then sent back to mixer **308** in preamplifier and mixer **305**. Mixer **308** mixes the output of filter **312** with the output of a local oscillator, frequency synthesizer **342**, which has a frequency chosen to be higher than the desired receiver carrier by 45.75 MHz. Thus, the output of mixer **308** is 45.75 MHz. There also is an image signal due to mixer **308** at 91.5 MHz above the input frequency, which is removed by filter **304** and filter **312**. Therefore, as the frequency of frequency synthesizer **342** is tuned to receive signals of different carrier frequencies, the bandpass and image reject filters **304** and **312** must also be tuned to properly pass only the desired signals and not the mixer images.

Frequency synthesizer **342** receives an input frequency reference signal (usually 16 bits) and outputs the status signals AUTOMATIC FREQUENCY CONTROL (AFC) ERROR and FREQUENCY (FREQ) LOCK. Additionally, a tuning signal which is used by the voltage controlled oscillator (VCO) in frequency synthesizer **342** is output from frequency synthesizer **342** to bandpass and image reject notch filters **304** and **312**. A local oscillator signal is output from frequency synthesizer **342** to mixer **308**.

The 45.75 MHz output signal of mixer **308** then passes through SAW (surface acoustic wave) filter **316**, which limits the bandwidth of the signal to only one (1) channel (6 MHz for NTSC standard) and applies a linear attenuation in frequency known as the Nyquist slope around the visual carrier frequency. The linear attenuation by SAW filter **316** converts the signal from a vestigial sideband signal to one which is equivalent to a single sideband with a carriers, so that the frequency response of the signal after demodulation is flat over the video bandwidth. SAW filter **316** is very "lossy" (on the order of 25 dB), so the input to SAW filter **316** is amplified by a preamplifier (not shown) by a corresponding amount to minimize noise effects.

The output of SAW filter **316** is input to intermediate frequency (IF) amplifier **320** in IF and baseband signal processor **310**. IF amplifier **320** provides most of the overall gain of TV tuner **300** and receives gain control from AGC **338**.

The output of IF amplifier **320** is sent to video detector **322** and is also sent off-chip to external video carrier filter **324**. This is the stage at which video demodulation is performed. Video detector **322** is essentially a mixer with the local oscillator input connected to the output of video carrier filter **324** through carrier amplitude limiter **326**. The output of the carrier limiter **326** is an in-phase representation of the video carrier signal without any modulation applied to it. The output of carrier limiter **326** is received by video detector **322**, which mixes the output of carrier limiter **326** with the output of IF amplifier **320**.

AFC frequency discriminator **340** is used in the prior art device to detect the difference between the carrier frequency contained in the output of carrier limiter **326** and a known valid carrier frequency reference. The output signal on the output of AFC frequency discriminator **340** is an error signal which is used to drive frequency synthesizer **342** in a direction that will reduce the error between the output of carrier limiter **326** and the known valid carrier frequency reference.

mechanically aligned during manufacture and are subject to variation in performance due to age, temperature, humidity, vibration and power supply performance. This was a critical drawback of previous tuners that had to be eliminated because it is a source of tremendous error and distortion, as well as complexity.

Furthermore, it is advantageous to have the up-conversion performed on-chip, because eliminating the inputting or outputting of high frequency signals to or from the integrated circuit of the present invention will avoid drive capability problems associated with high frequency signals and noise coupling problems associated with integrated circuit external interconnections.

The present invention next performs a down-conversion from the higher frequency using an image rejection mixing scheme that provides a tightly controlled down-conversion with very little added distortion. This also further minimizes the on-chip filtering effects by minimizing the amount of filtering that must be done at high frequency. The result is that high frequency operations within the present invention are limited to the initial stage inside the chip and are performed at a very low impedance level in order to minimize distortion. After the down conversions all remaining filtering is performed in continuous time mode filters that are realizable in an integrated circuit.

Another technical advantage of the present invention is that no manual-tuning of the final design is required. There are no external components needed to perform adjustments, as there is in the prior art.

A further technical advantage of the present invention is that the output impedance is selectable, based on system requirements, thereby minimizing power dissipation for certain applications. The present invention can drive a coaxial cable with a low output impedance driver or, for board level applications, the present invention can be configured to drive a high output impedance. The present invention can also match a very precise load impedance defined by a single external element.

The present invention advantageously utilizes much less board space than previous designs (on the order of 5% to 10% of the prior art designs) and dissipates only about one-fifth of the power. The present invention also advantageously operates on a single voltage level, as opposed to two or three levels for previous designs.

A further technical advantage of the present invention is that the need for a metal enclosure is reduced integration, by itself, allows for sufficient shielding to meet interference standards.

The monolithic television (MTV) tuner embodied in the present invention is intended to replace the TV tuner modules presently used in most broadcast television receiver devices. The MTV tuner integrates all of the functional elements of a tuner, except for the quartz crystal frequency reference and power supply bypass capacitors. The level of integration of the present invention dramatically reduces the cost of the basic TV tuner and enhances its manufacturability and reliability.

The TV tuner of the present invention is controlled externally by a computer or controller via a digital serial bus interface (I^2C). A preferred embodiment of the present invention provides an antenna input capable of being connected directly to a standard coaxial cable, thereby allowing both antenna and cable television applications. Furthermore, the baseband video and audio outputs are designed for high impedance or low impedance applications. The high impedance mode is a driver for short interconnections on a printed circuit and gives the benefit of low operating power. The low impedance mode drives an industry standard studio cable interface requiring greater supply power.

A preferred embodiment of the present invention is designed to operate on frequencies used for both over-the-air broadcasts and cable television with National Television Standards Committee (NTSC) encoded video. Receiver sensitivity is set to be limited by the antenna noise temperature for VHF systems. The present invention also employs a wide-range automatic gain control (AGC).

The baseband video output of the present invention is leveled, or has minimal variation in video amplitude with respect to antenna RF signal level, and can be configured via software for either low impedance studio standard loads or for high impedance loads with lower operating current. Audio output is broadband composite to allow connection to an external MTS decoder. The audio output can be configured for low impedance studio standard loads or high impedance loads via software control.

Control is accomplished via an I^2C bus interface. The bias and control circuits in a preferred embodiment of the present invention contain internal registers which can be updated via the control bus for operating frequency, video and audio transmission standards such as NTSC, PAL, SECAM and MTS, power, and test modes. Status of the bias and control circuits can be checked via a status register accessible through the I^2C bus interface. Status data include AFC error, channel lock and received signal strength indicator.

The operating frequency of the present invention is referenced to an external crystal or reference frequency generator. A minimum of external components are used with the present invention and no tuning of any components is required.

The present invention may be implemented in Bipolar, BiCMOS, or CMOS processes, however, a preferred embodiment of the present invention employs a BiCMOS process to reduce the difficulty in developing the design by allowing maximum flexibility.

A preferred embodiment of the present invention performs sound processing after IF amplification and prior to video demodulation. This method is known as "quasi-split" sound and is not used in the majority of television receivers for cost reasons. Quasi-split sound does not have a "buzz" in the audio signal during periods when the video picture is very white. In the integrated circuit of the present invention, quasi-split sound is a negligibly small cost and thus may be integrated in all embodiments of the present invention, integration of quasi-split sound further reduces component count and does not require external manual adjustments.

The foregoing has outlined rather broadly the features and technical advantages of the present invention in order that the detailed description of the integrated television tuner that follows may be better understood. Additional features and advantages of the monolithic television tuner will be described hereinafter which form the subject of the claims of the invention. It should be appreciated by those skilled in the art that the conception and the specific embodiment disclosed may be readily utilized as a basis for modifying or designing other structures for carrying out the same purposes of the present invention. It should also be realized by those skilled in the art that such equivalent constructions do not depart from the spirit and scope of the invention as set forth in the appended claims.

BRIEF DESCRIPTION OF THE DRAWINGS

For a more complete understanding of the present invention, and the advantages thereof, reference is now

and the IF output is given by:

$$V_{IF} = V_{RF} \cdot V_{LO} \quad [3]$$
$$= AB/2[\cos(\omega_{RF} - \omega_{LO})t + \cos(\omega_{RF} + \omega_{LO})t].$$

In the frequency domain, the IF frequencies are given by: $f_{IF}=|f_{RF}\pm f_{LO}|$. In realizable mixers, there are various non-linearities which cause the generation of harmonics at multiples of the RF and IF frequencies. Thus, the IF spectrum contains harmonics at $f_{IF}=mf_{RF}\pm nf_{LO}$, where m and n are integers.

For example, if the output of local oscillator **450** is 400 MHz and the desired input signal to be demodulated from the antenna is 800 MHz, a basic mixer would perform an additive mix to obtain a 1200 MHz output at the output of mixer **408**. However, since frequency synthesizer **450** also has a harmonic at 800 MHz, and a signal exists on the antenna at 400 MHz, these signals are also added to produce an 800 MHz output that would interfere with the desired signal to be demodulated at 800 MHz.

Mixer **408** is a subtractive mixer that will subtract the frequency of local oscillator **450** from the input RF spectrum between 0 Hz to 900 MHz that is filtered by RFLPF **404**. For example, if the desired frequency to be demodulated in the RF input is 400 MHz, the local oscillator would be set to 1600 MHz and a basic mixer would generate two IF outputs:

$$f_{IF}=|400-1600|=1200 \text{ MHz and} \quad [4]$$

$$f_{IF}=|400+1600|=2000 \text{ MHz} \quad [5]$$

Since mixer **408** is a subtractive mixers the 2000 MHz signal will be filtered out and only the 1200 MHz signal will be passed on to mixer **410**. Also, since mixer **408** is a subtractive mixer, the incoming spectrum is inverted on the output, so than the video carrier is 1.25 MHz below the upper edge of the 6 MHz channel bandwidth, the color carrier is 3.58 MHz below the video carrier and the aural carrier is 4.5 MHz below the video carrier. For example, the 6 MHz bandwidth of channel **2** from 54–60 MHz would appear inverted between 1201.25 MHz and 1195.25 MHz. The channel **2** video carrier would appear at 1200 MHz, the color carrier would appear at 1196.42 and the aural carrier would appear at 1195.5 MHz.

As noted above, it is ms a characteristic of both the input RF signal received, by mixer **408** from LNTA **406** and of frequency synthesizer **405** that harmonic signals exist at multiples of the desired frequencies. By using a subtractive mixer and a sufficiently high-frequency output from oscillator **450**, the harmonics on the output of frequency synthesizer **450** will all be more than 1200 MHz higher in frequency than any of the input frequencies existing in the 0 Hz to 900 MHz range from antenna **402**.

For example, the lowest output of local oscillator 450 is 1200 MHz. The first harmonic of this signal exists an 2400 MHz. Since the highest output signal passing through LNTA **406** is 900 MHz, the subtractive mix of the 2400 MHz harmonic and the 900 MHz signal from antenna **402** would yield a signal of |900–1200|=1300 MHz. This is higher than the desired 1200 MHz output from mixer **408**, and may therefore be filtered out by low-pass filtering. Using a local oscillator whose output is higher that the RF input frequency is known as high-side injection.

The first IF signal of 1200 MHz is mixed in second mixer (MIX2) **410**, which is an image-rejection mixer, with the fixed 1180 MHz reference output of second local oscillator (LO2) **412** to generate the second IF at 20 MHz visual carrier. Because the RF input signal is lower in frequency than the LO referenced, the mixing of the two signals will result in a down conversion of the RF input. The IF frequencies, given by $f_{IF}=|f_{RF}\pm f_{LO}|$, have carriers at f_{IF}= 1200–1180=20 MHZ and at f_{IF}=1200+1180=2380 MHZ. The image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at f_{IF}=|1160–1180|=20 MHz.

The image rejection mixer receives an input signal from MIX1 that has the general form:

$$V_{RF}=A\cos(\omega_{RF}t), \quad [6]$$

and the local oscillator signal from LO2 is split into two phase quadrature signals having the general form:

$$V_{LOI}=B\cos(\omega_{LO}t) \text{ and} \quad [7]$$

$$V_{LOQ}=B\sin(\omega_{LO}t). \quad [8]$$

The desired output signal, V_{IF}, resulting from the mixing of the RF input and the phase quadrature signals has the general form:

$$V_{IF}=AB\cos(\omega_{RF}-\omega_{LO})t. \quad [9]$$

The desired mixer signal at $\omega_{IF}=\omega_{RF}-\omega_{LO}$ appears in the mixer outputs while the undesired mixer image signal at $\omega_{IF}=\omega_{RF}+\omega_{LO}$ (−1160+1180=20) is rejected.

FIG. 5 depicts an alternate embodiment of the RF front-end (i.e., RF converter **110** and IF processor **120**) of the present invention. First IF filter (FIFF) **509** between mixer **408** and mixer **510** filters out all frequency components of the image falling on 1160 MHz, such that mixer **510** is a basic mixer similar to mixer **408**, rather than an image rejection mixer. While the image rejection mixer is easier to integrate, it has accuracy limitations, and noise and distortion figures may suffer.

Returning to FIG. 4, the second IF signal is next filtered for the final video bandwidth by second IF filter (SIFF) **420**, a low-distortion continuous-time bandpass filter. SIFF **420** also comprises an automatically tuned low-pass Nyquist slope filter (NSF) for VSB balancing. The −6 dB point on the slope of the NSF is maintained at the visual carrier frequency (20 MHz).

The output of SIFF **420** is coupled to IF amplifier (IFAMP) **422**, a gain-controlled amplifier which applies up to 80 dB of gain. The output of IFAMP **422** is then mixed in video detector (VDET) **426** with the video carrier, which has been extracted by video extraction filter (VCEF) **425** and video carrier limiter (VCLIM) **428** to produce the baseband video output.

The video carrier is compared by automatic frequency control (AFC, frequency detector (DET) **440** with a synthesized 20 MHz reference signal from AFCREF **442** to generate a frequency error signal which is passed back to LO1 for automatic frequency control.

The baseband video from VDET **426** is passed through a low-pass video bandwidth filter (VBBF) **430** to remove detection harmonics. The synchronization pulses are clamped to a reference level by sync clamp (SCMP) **432**, which contains a phase-locked loop that is locked to the horizontal sync rate, RX LOCK **434**, or vertical rate as an option, to control the clamp timing. Timing is also generated in SCMP **432** for AGC **438** and NINV **436**. SCMP **432** amplifies the video to its output voltage level by a fixed gain of 20 dB. The clamped video is processed by noise inverter (NINV) **436** to remove excessively large positive or negative amplitude noise pulses.

The resulting video is tested for blanking to sync level by AGC **438**, which adjusts the gain of LNTA **406** and IFAMP

The output of the video detector **322** is a baseband video signal combined with several high frequency mixing artifacts. These artifacts are removed by a video baseband filter **330**. The output of video baseband filter **330** is fed to synchronization pulse clamp (sync clamp) **332**, which sets the level of the sync pulses to a standard level.

Next, the output of sync clamp **332** is sent to noise inverter **334**, which removes large noise spikes from the signal. The output of noise inverter **334** is sent to video buffer **336**, which is configured to drive fairly high circuit board impedances of approximately 1000 to 2000 ohms.

The output of noise inverter **334** is also sent to AGC (automatic gain control) **338**, which compares the level of the synchronization pulses to the signal blanking level to measure the incoming signal strength and generates a gain control signal which is used by IF amplifier **320** and RF preamplifier **306** to dynamically adjust the gain of the TV tuner **300** for the correct level at the final output.

The audio signal is an FM signal which follows the same path as the video through video detector **322**. At the output of video detector **322**, the audio signal appears as a subcarrier at 4.5 MHz, due to the fact that the audio signal comes into prior art TV tuner **300** 4.5 MHz higher in frequency than the desired video carrier. The audio subcarrier is passed on to an FM quadrature demodulator. The FM quadrature demodulator is comprised of a mixer, audio second detector **350**, and a 90 degree (at 4.5 MHz) phase shifter, audio carrier phase shifter **360**. The output of the audio second detector **350** is a baseband audio signal, which is filtered by lowpass (30 kHz) filter **352** to remove any undesired high frequency components. The output of lowpass filter **352** is finally passed on to audio buffer **354**, which drives an audio amplifier that ultimately drives a speaker. Serial digital interface **344** receives SERIAL DATA and SERIAL CLOCK inputs to provide control and update status for the prior art television receiver.

Baseband and image reject notch filters **304** and **312** are typically comprised of a plurality of capacitors, inductors and varactor diodes. Video carrier filter **324** is usually comprised of three discrete elements: an inductor and two capacitors. Likewise, audio carrier phase shifter **360** is also comprised of an inductor and two capacitors. In addition to the circuit elements shown as discrete components outside of circuit elements **305**, **310** and **315** in FIG. **3**, other discrete components (not shown) are connected to IF and baseband signal processor **310** and frequency synthesizer **342** for tuning purposes. Frequency synthesizer **342** is typically tuned by several external capacitors, inductors and/or varactor diodes. Video buffer **336** and audio buffer **354** will also typically employ external discrete elements, such as resistors, capacitors and/or transistors. Video baseband filter **330** and low pass filter **352** may also employ external inductors and capacitors.

DETAILED DESCRIPTION OF THE INVENTION

FIG. **1** depicts a high-level block diagram of the functional elements contained in a preferred embodiment of TV tuner **100**. The RF signal received from the antenna or other source is converted by RF converter **110** to an intermediate frequency (IF) signal and sent to IF processor **120**. The output of IF processor **120** is sent to Vestigial Sideband (VSB) detector **130** and video processor and driver **150** to produce the video output signal of the present invention. The video signal is detected via a synchronous AM demodulator.

The output of IF processor **120** is also sent to FM detector **140** and audio processor and driver **160** in order to produce

the audio output signal of the present invention. Audio detection is performed by a quadrature FM detector utilizing a phase-locked loop for the quadrature reference. I^2C interface **170** receives I^2C DATA and I^2C CLOCK signals from the I^2C interface bus. Frequency reference **190** is an oscillator that is synchronized to the external reference crystal. Bias and control logic circuit **180** defines internal bias voltages and currents and maintains the status and control registers of the television tuner of the present invention.

FIG. **2** illustrates a pin layout diagram of a preferred embodiment of an integrated circuit in accordance with the present invention. Integrated circuit **200** is connected to a supply voltage on pins **11** and **20** (VCC1 and VCC2) and is grounded on pins **3**, **4**, **6–10**, **14** and **17** (GND1–GND9). An input reference signal from an external crystal oscillator is supplied on pins **1** and **2** (CRYSTAL 1 and CRYSTAL 2/REF IN). Integrated circuit **200** is connected to an IIC (I^2C) interface bus on pin **16** (SERIAL CLOCK) and pin **15** (SERIAL DATA). The RF input from an antenna or other source is connected to pin **5** (RF INPUT). Integrated circuit **200** outputs video and audio signals on pins **18** and **19** (VIDEO and VIDEO REF) and pins **12** and **13** (AUDIO and AUDIO REF).

FIG. **4** is a detailed electrical block diagram of TV tuner **100** in accordance with a preferred embodiment of the present invention. FIG. **4** depicts an up-converting dual conversion superheterodyne receiver. The Vestigial Sideband (VSB) encoded video signal is processed via a Nyquist slope receiver attenuation characteristic filter prior to detection.

The RF signal enters TV tuner **100** from the antenna **402** (or cable, not shown) and is passed through RF low-pass filter (RFLPF) **404** to limit the incoming band to below 900 MHz. The filtered RF signal is amplified up to 20 dB by a gain-controlled low-noise transconductance amplifier (LNTA) **406**.

The input signal received by antenna **402**, filtered by RFLPF **404** and amplified by LNTA **406** is the standard broadcast television spectrum. Channels **2–13** in the VHF band extend from 54 Megahertz to 216 MHz and channels **14–83** in the UHF band extend from 410 MHz to 890 MHz. Each channel in the VHF and UHF band has a 6 MHZ bandwidth with the video carrier frequency located 1.25 MHz above the lower band edges a color carrier frequency located 3.58 MHz above the video carrier, and an audio carrier frequency located 4.5 MHz above the video carrier. For example, Channel **2** has a 6 MHz bandwidth from 54 to 60 MHz, a video carrier at 55.25 MHz, a color subcarrier at 58.83 MHz and an audio carrier at 59.75 MHz.

The output of first local oscillator (LO1) **450**, operating between 1200 and 2100 MHz, is mixed in first mixer (MIX1) **408** with the RF signal to generate a first IF video carrier frequency of 1200 MHz. This approach leads to minimum distortion due to mixer images and harmonic mixing. The first IF is crudely filtered by the bandwidth limitation of first mixer **408** to minimize harmonic effects.

In a basic mixer scheme, a mixer receives two inputs, an RF input and a local oscillator (LO) input, and generates an IF output. The RF input has the general form:

$$V_{RF}=A\cos(\omega_{RF}t), \quad [1]$$

the local oscillator input has the general form:

$$V_{LO}=B\cos(\omega_{LO}t), \quad [2]$$

just an ordinary subtractive mixer. An important feature of MIX2 is that it must achieve a high image rejection (at least 50 dB) without the need for post-fabrication trimming techniques to calibrate MIX2. As shown in FIG. 6, the 50 dB image rejection may be achieved by paralleling two Gilbert cell multipliers **610** and **612** at MIX2 input and using fairly high-order balanced RC-phase shifters **613**, **614** with very low component sensitivity with resulting phase errors via summer **615** of less than 0.2 degrees over a 20 MHz bandwidth. Multipliers **610** and **612** receive inputs from Phase Splitter **611** and MIX1 **408**. These are traditionally avoided due to relatively high noise levels, however, in a preferred embodiment of TV tuner **100**, the preious stages provide sufficient gain reduce the noise impact.

TABLE 4

MIX2 Specifications

Parameter	Description	Min	Typ	Max	Unit
G_I	Current Gain		0		dB
NF	Noise Figure			10	dB
f_{RF}	RF Input Frequency	1195		1205	MHz
f_{LO}	LO2 Input Frequency		1180		MHz
f_{IF}	IF Output Frequency	15.25		21.25	MHz
V_{LO}	LO2 Input Valtage		0.2		V(rms)
A_{IM}	Input Image Rejection	60			dB

Second IF Filter (SIFF)

Table 5 lists the operating parameters of second IF filter (SIFF) **420** in accordance with a preferred embodiment of the present invention. The 20 MHz IF signal is filtered for channel bandwidth in this integrated transconductor-capacitor (G_m-C) filter. A transitional Gaussian bandpass filter is used to provide adjacent channel and noise rejection. SIFF **420** keeps the distortions due to noise and out-of-band signals below −54 dBc. The in-channel group delay is to be maximally flat to the color subcarrier frequency. The aural signal passes through SIFF **420**. Since the up-conversion scheme inverts the channel frequencies at the output of MIX2, the visual carrier is now at 20 MHz and the aural carrier is at 15.5 MHz. The adjacent visual carrier is now at 14 MHz and the adjacent aural carrier is at 21.5 MHz leading to the filter specifications given below.

TABLE 5

SIFF Specifications

Parameter	Description	Min	Typ	Max	Unit
f_{pl}	Lower Passband Limit $A(\omega) < 0.5$ dB		16.75		MHz
f_{ph}	Upper Passband Limit $A(\omega) < 0.5$ dB		19.25		MHz
f_{sl}	Lower Stopband Limit $A(\omega) > 54$ dB		15.25		MHz
f_{sh}	Upper Stopband Limit $A(\omega) > 54$ dB		20.75		MHz
G_V	Voltage Gain		0		dB
IP_3	Third Order Input Intercept	20			dBm
NF	Noise FIgure			15	dB

IF Amplifier (IFAMP)

Table 6 lists the operating parameters of IF amplifier (IFAMP) **422** in accordance with a preferred embodiment of the present invention. The 20 MHz signal is amplified by 80 dB in this block. The gain can be reduced 0 dB by application of an automatic gain control (AGC) signal voltage. The AGC is implemented in fully differential form for best on-chip noise rejection.

TABLE 6

IFAMP Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain - No AGC		80		dB
G_V	Voltage Gain - Full AGC		20		dB
NF	Noise Figure			10	dB
BW	Bandwidth	25			MHz

Visual Carrier Extraction Filter (VCEF)

Table 7 lists the operating parameters of visual carrier extraction filter (VCEF) **425** in accordance with a preferred embodiment of the present invention. The video carrier at 15 MHz is extracted from IFAMP **422** output for synchronous detection processing. VCEF **425** passes along any incidental carrier phase modulation (ICPM) in the Nyquist filter without any group delay distortion for best detector performance.

TABLE 7

VCEF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_0	Center Frequency		20		MHz
BW	Bandwidth	0.5	1	3	MHz
NF	Noise Figure			5	dB

Visual Carrier Limiter (VCLIM)

Table 8 lists the operating parameters of visual carrier limiter (VCLIM) **428** in accordance with a preferred embodiment of the present invention. The synchronous detector requires a fixed amplitude reference carrier. This is derived via VCEF **425** and then passed through VCLIM **428**. With input level variations of 20 dB and depth of modulation possibly exceeding 95 percent, VCLIM **428** output is stable within an accuracy of −50 dB. The circuit delay of VCEF **425** and VCLIM **428** combined must not exceed 45 degrees at the visual carrier frequency so that the detector output level and noise behavior is not degraded.

TABLE 8

VCLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	46			dB
V_{th}	Limiter Input Threshold		10		mV(p—p)
V_{limit}	Limiter Output Level	2			V(p—p)
NF	Noise Figure			5	dB

Video Detector (VDET)

Table 9 lists the operating parameters of video detector (VDET) **426** in accordance with a preferred embodiment of the present invention. Video is detected synchronously by applying the output of IFAMP **422** as a first input signal and

422 to maintain a standard 1 Volt peak-to-peak video signal. The video at this stage is buffered by video buffer (VBUF) 437 for off-chip drive using one of three signal impedance standards. VBUF 437 can also mute the signal.

Audio processing is accomplished by extracting the audio signal at the output of IFAMP 422 through narrow bandpass audio IF filter (AIFF) 427. AIFF 427 produces an FM signal that is limited by audio IF limiter (AIFLIM) 444 to remove AM noise and distortion.

A delay-locked loop comprised of audio phase detector (APD) 455, audio delay-locked loop low-pass filter (ADLPF) 458, and audio phase shifter (APS) 460 generates a signal that is delayed by 90° at the FM carrier frequency and mixes that signal with the limited FM signal in audio quadrature detector (AQD) 462 for quadrature detection of the FM. The audio baseband is passed through a 120 kHz low-pass audio baseband filter (ABF) 464 to remove detection artifacts. The audio is buffered in audio buffer (ABUF) 466 for off-chip drive with selectable drive impedances at a 0.35 Volt (rms) level.

Control registers 471 and status registers 472 are interfaced to serial digital interface 170. Serial digital interface 170 in turn receives external commands from an external microprocessor or microcontroller through SERIAL DATA and SERIAL CLOCK control lines. In a preferred embodiment of the present invention, serial digital interface 170 may be a Inter integrated Circuit (IIC or I^2C) interface, which is a proprietary specification of Philips Corporation.

System Block Specifications

The overall design of TV tuner 100 is guided by the limitations of what can be reliably designed on an integrated circuit. Blocks of TV tuner 100 are specified by referring to actual circuit elements so that capabilities such as noise, maximum signal handling, intermodulation and gain can be ascertained. Perhaps, the most important element of such a design is to limit expected chip performance so that it can be manufactured with a high yield and, therefore, a low cost. Note that all measurements in dBm are referenced to a standard video RF and baseband impedance of 75 Ω.

RF Low-Pass Filter (RFLPF)

Table 1 lists the operating parameters of RF low-pass filter (RFLPF) 404 in accordance with a preferred embodiment of the present invention. RFLPF 404 is a 900 MHz third order low-pass filter comprising package lead inductance, printed circuit capacitance and on-chip capacitors. For most applications RFLPF 404 should be satisfactory for image (2400–3300 MHz) rejection in MIX1. External filtering matched to 75 ohm impedance could be added for extreme situations, such as very large out-of-band signals.

TABLE 1

RFLPF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	−1		0	dB
f_c	Cutoff Frequency (−3 dB)		900		MHz

Low Noise Transconductance Amplifier (LNTA)

Table 2 lists the operating parameters of low-noise transconductance amplifier (LNTA) 406 in accordance with a preferred embodiment of the present invention. LNTA 406 is the front end of the receiver. The single-ended input signal is converted to a differential current which will be passed through a mixer chain. The critical function of this circuit is to maintain a very low noise figure (NF), a substantial gain (G_V), and a high third-order intermodulation intercept (IP_3). The gain is determined relative to the rest of the circuit so that the noise figure of this block dominates the total receiver noise figure. Note that, since this is a transconductance amplifier, the gain is defined by the load impedance of the final mixer stage and must be properly defined before a final determination of LNTA 406 performance. AGC capability is included to reduce the impact of distortion, such as intermodulation or crossmodulation, in the presence of high level signals. The input is impedance matched to 75 Ω.

TABLE 2

LNTA Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain - no AGC	20		30	dB
G_{VAGC}	Voltage Gain - full AGC	−10		0	dB
NF	Noise Figure	4.5		5.4	dB
P_{-1dB}	1 dB Gain Compression Input Level	−24			dBm
IP_3	Third Order Input Intercept	−10			dBm
f_{in}	Input Frequency	40		900	MHz
I_{DC}	DC Bias Current		20		mA

First Mixer (MIX1)

Table 3 lists the operating parameters of first mixer (MIX1) 408 in accordance with a preferred embodiment of the present invention. Proper rejection of any undesired images and spurious signals due to local oscillator harmonics requires that MIX1 be configured for an up-conversion scheme with high-side injection. Thus, the output of MIX1 for the visual carrier is at the microwave frequency of 1200 MHz, the audio carrier is at 1195.5 MHz, and local oscillator 450 input runs between 1.2 and 2.1 GHz. A single-pole low-pass element is designed into the output circuitry of MIX1 to assist in the rejection of any down-converted signals due to local oscillator harmonics. The noise and distortion effects of MIX1 are included in LNTA 406 specifications. The design will include a bond-wire parallel LC network option to further band-limit any image energy at the output of this mixer if necessary.

TABLE 3

MIX1 Specifications

Parameter	Description	Min	Typ	Max	Unit
f_{RF}	RF Input Frequency	40		900	MHz
f_{LO}	LO1 Input Frequency	1200		2100	MHz
f_{IF}	IF Output Frequency	1195		1205	MHz
V_{LO}	LO1 Input Voltage		0.2		V(rms)

Second Mixer (MIX2)

Table 4 lists the operating parameters of second mixer (MIX2) 410 in accordance with a preferred embodiment of the present invention. The first IF of 1.2 GHz (visual carrier) is input to MIX2, which is an image-rejection mixer, which down-converts to a 20 MHz second IF. The image reaction is done to negate the need for filtering of the first mixer output in order to reject signals falling on the image at 1160 MHz from being down-converted by MIX2 if MIX2 were

transient response time to account for aircraft induced flutter (fading) and unconditional stability. The response can be changed to update gain only during the vertical sync period, just as with SCMP 432, so that scrambled video may pass through the tuner without any adverse effects.

The AGC voltages defined by the performance of the RF and IF amplification behavior and input signal level, is measured via 4 bit analog-to-digital converter and stored as the data word Received Signal Strength Indicator (RSSI) 439. Automatic gain control is applied first to IFAMP 422 to reduce the video level, and beyond 50 dB of reduction, also to LNTA 406, which implements a delayed AGC function.

TABLE 13

AGC Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p—p)
V_{s-b}	Input Sync to Blanking Level Reference		0.286		V
G_{AGC}	AGC Loop Gain	70			dB
GBW_H	AGC Gain-Bandwidth H-Mode		150		Hz
GBW_V	AGC Gain-Bandwidth V-Mode		6		Hz
ϕ_{MAGC}	AGC Loop Phase Margin	60			°

Video Output Buffer (VBUF)

Table 14 lists the operating parameters of video buffer (VBUF) 437 in accordance with a preferred embodiment of the present invention. The video is at an internal level of 1 volt peak-to-peak differential. VBUF 437 converts this signal to a current via a linear transconductance amplifier that drives a load resistor. Feedback is arranged via an external signal to allow for continuous adjustment of the gain and offset of the amplifier so that no AC coupling is required in the next stage. An on-chip integrator is used to define the feedback level and reject the video signal. The control signal VOUT MODE 445 enables on-chip loads of 75 Ω, 1000 Ω or no load, which equates to a current source output. The control signal VIDEO MUTE 446 enables a video mute switch which sets the output to the blanking level.

TABLE 14

VBUF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p—p)
V_{OV}	Video Output Level (R_L > 70 Ω)		1		V(p—p)
V_{OVR}	Video Output DC Reference (R_L > 70Ω)	0.5		3.5	V
Z_{OUT}	Output Impedance 75Ω Mode	73	75	77	Ω
Z_{OUT}	Output Impedance 1000Ω Mode	980	1000	1020	Ω
Z_{OUT}	Output Impedance Current Mode		10		kΩ

Automatic Frequency Control (AFC) Frequency Detector

Table 15 lists the operating parameters of automatic frequency control (AFC) frequency detector (DET) 440 in accordance with a preferred embodiment of the present invention. The extracted IF video carrier is compared to a fixed reference at 20 MHz and used to drive LO1 in a direction to equalize the two. The error in LO1 from its frequency control setting is the signal AFC ERROR 452. The 20 MHz reference is generated via a frequency synthesizer from the master reference. The 20 MHz reference is compared to the extracted video carrier signal at the output of VCEF 425 using a digital frequency comparator which drives a seven bit up-down counter that is summed with the low bits of the frequency code sent to LO1 from the I²C interface 170. Thus, the frequency error appears in the counter which can be read via the interface bus. This counter is set to 1000000b when there is no frequency error. This condition sets the status signal FREQ LOCK 453.

TABLE 15

AFC Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{CIN}	Video Carrier Input Level		0.1		V(p—p)
f_{ERROR}	Frequency Error From Nominal	−500		500	kHz
f_{UPDATE}	Frequency Update Rate			30	Hz

AFC Reference Oscillator (AFCREF)

Table 16 lists the operating parameters of AFC reference oscillator (AFCREF) 442 in accordance with a preferred embodiment of the present invention. AFCREF 442 is a fixed frequency synthesizer running at 20 MHz, which is phase-locked to the quartz crystal frequency reference. Phase noise is not particularly critical due to the application of this circuit at low modulation frequencies only.

TABLE 16

AFCREF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p—p)
N_ϕ	Phase Noise for 20 Hz < f_m < 1 MHz			−60	dBc
f_0	Operating Frequency		20		MHz

First Local Oscillator (LO1)

Table 17 lists the operating parameters of first local oscillator (LO1) 450 in accordance with a preferred embodiment of the present invention. LO1 is a phase-locked frequency synthesizer configured for specific levels of phase noise to allow for best receiver sensitivity without excessive spurious noise responses. It is digitally controlled and utilizes a quartz crystal reference for frequency and in-band phase noise control. The frequency control input FREQUENCY 451 from the serial digital interface 170 bus has sixteen (16) bits with 62.5 kHz LSB resolution. Four (4) more bits below the LSB as well as the first 3 LSB's are used by the AFC for fine tuning.

TABLE 17

LO1 Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p—p)
N_ϕ	Phase Noise for f_m > 1 MHz			−110	dBc
N_ϕ	Phase Noise for 60 Hz < f_m < 1 MHz			−80	dBc

the output of VCLIM **428** as a second input signal to a double-balanced mixer (VDET **426**). The output of VDET **426** is configured with a single low-pass pole set to allow 4.2 MHz to pass with a 0.5 dB loss.

TABLE 9

VDET Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
BW	Output −0.5 dB Bandwidth		4.2		MHz
V_{video}	Video Output Level		0.1		V(p—p)
NF	Noise Figure			5	dB

Video Baseband Filter (VBBF)

Table 10 lists the operating parameters of video baseband filter (VBBF) **430** in accordance with a preferred embodiment of the present invention. The detected video is band-limited to 4.2 MHz by a multiple pole transitional Gaussian response-low-pass filter which has been predistorted for both compensation of SIFF **420** response and group delay characteristics conforming to FCC regulations. VBBF **430** rejects the video carrier at 20 MHz and the high level 40 MHz modulated signal at the output of VDET **426** with minimum pass-band transient distortion.

TABLE 10

VBBF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_p	Passband Corner Frequency A(ω) < 0.5 dB		4.2		MHz
f_s	Stopband Corner Frequency A(ω) > 54 dB		8.0		MHz
NF	Noise Figure			5	dB

Sync Clamp (SCMP)

Table 11 lists the operating parameters of sync clamp (SCMP) **432** in accordance with a preferred embodiment of the present invention. The sync pulses are clamped via two techniques. A diode clamp is used to define the minimum level to which the sync tip is set. Once this is achieved, after one sync pulse, a sync slicer regenerates the sync pulses only and passes these on to a phase-locked loop (PLL). The PLL acquires the horizontal sync rate and generates a sync gating pulse approximately centered in the horizontal sync pulse.

Once the PLL has achieved lock, a gated clamp circuit using the sync gating pulse is enabled to define the sync level and the diode clamp is disabled. For processing of scrambled video, a line counter is inserted into the phase-locked loop to count a field, the sync is integrated for detection of vertical sync and only the vertical sync pulse is used for the clamp gate. The sync gating pulse is used, via a clocked delay of 5 microseconds to generate a blanking gate used for AGC processing. SCMP **432** also amplifies the video to its final output level. The output signal RX LOCK **434** becomes active when the phase-locked loop has achieved lock to the sync pulses.

TABLE 11

SCMP Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		20		dB
V_{Video}	Video Output Level		1		V(p—p)
f_{PLLH}	PLL Operating Frequency H-Sync Mode		15750		kHz
f_{PLLV}	PLL Operating Frequency V-Sync Mode		60		Hz

Noise Inverter (NINV)

Table 12 lists the operating parameters of noise inverter (NINV) **436** in accordance with a preferred embodiment of the present invention. SCMP **432** video output will include extreme transient noise pulses in both black and white directions. The black noise is typically a large RF transient due to some form of interference. The signal will drop below the blanking level during the horizontal line picture period and may upset downstream sync detect circuits. A comparator detects these levels below the blanking level during active picture, or non-blanked, periods and enables an inverting video path during the event which is summed with the original video. The summation cancels the effect of the noise pulse so that the sync level region is not disturbed. Extreme positive excursions, which appear as white, are also cancelled when they exceed the equivalent **108** IRE white level, or less than 5 percent modulation. These typically are due to desensitization effects of the receiver when pulse interference exists on an out of band frequency and typically result in the disruption of the synchronous detector. Positive noise inversion is accomplished using a comparator referenced to the 108 IRE level and enabling a video inversion path to a summer as for the negative case above.

TABLE 12

NINV Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p—p)
V_{WINS}	White Noise Inverter Insertion Level		95		IRE
V_{BINS}	Black Noise Inverter Insertion Level		10		IRE
V_{WTH}	White Inverter Threshold		108		IRE
V_{BTH}	Black Inverter Threshold Non-Blanked		0		IRE

Automatic Gain Control (AGC)

Table 13 lists the operating parameters of automatic gain control (AGC) **438** in accordance with a preferred embodiment of the present invention. The clamped video is fed into a sample-and-hold circuit controlled by the blanking gate. The sample-and-hold circuit output drives a differential amplifier which has a reference level at its other input equal to the desired output blanking level for the standard NTSC video waveform.

On each horizontal sync pulse the output of the differential amplifier represents the gain error of the receiver to achieve the video waveform. This is fed back to LNTA **406** and IFAMP **422** to adjust the gain as necessary. The timing and gain of the differential amplifier must be set for fast

embodiment of the present invention. AQD **462** is a mixer that is used to compare the incoming audio IF signal with a delayed version of the IF signal running in quadrature with the incoming carrier. The output is the desired audio baseband signal.

TABLE 24

AQD Specifications

Parameter	Description	Min	Typ	Max	Unit
G_v	Voltage Gain		0		dB

Audio Baseband Filter (ABF)

Table 25 lists the operating parameters of audio baseband filter (ABF) **464** in accordance with a preferred embodiment of the present invention. ABF **464** is a high-order Chebychev low-pass design with a cutoff at 120 kHz. This is wide enough to pass all multiplex audio subcarriers for external decoding.

TABLE 25

ABF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		10		dB
f_p	Passband Corner Frequency $A(\omega)$ <0.5 dB		120		kHz
f_s	Stopband Corner Frequency $A(\omega)$ >54 dB		300		kHz

Audio Output Buffer (ABUF)

Table 26 lists the operating parameters of audio output buffer (ABUF) **466** in accordance with a preferred embodiment of the present invention. The audio is at an internal level of 0.1 volts peak to peak. ABUF **466** converts this signal to a current via a linear transconductance amplifier which then drives a load resistor. Feedback is arranged via an external signal to allow for continuous adjustment of the gain and offset of the amplifier so that no AC coupling is required to the next stage. An on-chip integrator is used to define the feedback level and reject the audio signal.

The control signal AOUT MODE **467** enables on-chip loads of 600 Ω, 10 kΩ or no load, which equates to a current source output. The control signal AUDIO MUTE **468** controls an audio mute switch which disables the output.

TABLE 26

ABUF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Audio Input Level		0.1		V (p-p)
V_{OV}	Audio Output Level (R_L >500Ω)		0.35		V (p-p)
V_{OVR}	Audio Output DC Reference (R_L > 500Ω)	0.5		3.5	V
Z_{OUT}	Output Impedance 600Ω Mode	588	600	612	Ω
Z_{OUT}	Output Impedance 10 kΩ Mode	9.8	10.0	10.2	kΩ
Z_{OUT}	Output Impedance Current Mode		100		kΩ

Serial Digital Interface

Tables 27–29 define the registers, register bits and bit functions that are addressable by serial digital interface **170** in accordance with a preferred embodiment of the present invention. For example, control and status may be performed via an inter integrated Circuit (I^2) bus interface. Serial digital interface **170** contains all registers and provides access to all parallel digital on-chip functions via the serial bus.

TABLE 27

Register Definition

Register Name	Description	Register Address	Read/Write
FREQ0	Frequency - Low byte	00	Read/Write
FREQ1	Frequency - High byte	01	Read/Write
CTRL0	Functional control	02	Read/Write
CTRL1	Functional control	03	Read/Write
STAT0	Functional status	04	Read
STAT1	Functional status	05	Read
STAT2	Functional status	06	Read
CMPY0	Company code - Low byte	07	Read
CMPY1	Company code - High byte	08	Read
REV0	Revision code - Low byte	09	Read
REV1	Revision code - High byte	0A	Read

The definitions of the bits used in each register are.

TABLE 28

Register Bit Definitions

Register	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2	Bit 1	Bit 0
FREQ0	F7	F6	F5	F4	F3	F2	F1	F0
FREQ1	F15	F14	F13	F12	F11	F10	F9	F8
CTRL0	PWRDN	AMUTE	VMUTE		MODE3	MODE2	MODE1	MODE0
CTRL1					VRS1	VRS0	ARS1	ARS0
STAT0	RLOCK	FLOCK				AFC2	AFC1	AFC0
STAT1					RSSI3	RSSI2	RSSI1	RSSI0
STAT2	T7	T6	T5	T4	T3	T2	T1	T0
CMPY0	C7	C6	C5	C4	C3	C2	C1	C0
CMPY1	C15	C14	C13	C12	C11	C10	C9	C8
REV0	R7	R6	R5	R4	R3	R2	R1	R0
REV1	R15	R14	R13	R12	R11	R10	R9	R8

The status registers are read only. Attempting to write a status register will have no effect. The function of each bit is defined as:

TABLE 17-continued

LO1 Specifications

Parameter	Description	Min	Typ	Max	Unit
f_0	Operating Frequency	1200		2100	MHz
f_{STEP}	Frequency Step		3.906		kHz
f_{REF}	Reference Frequency		62.5		kHz

Second Local Oscillator (LO2)

Table 18 lists the operating parameters of second local oscillator (LO2) **412** in accordance with a preferred embodiment of the present invention. LO2 is a fixed tuned frequency synthesizer set to 1180 MHz. It uses the same frequency reference as LO1 and has similar phase noise characteristics.

TABLE 18

LO2 Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p—p)
N_ϕ	Phase Noise for $f_m > 1$ MHz			−110	dBc
N_ϕ	Phase Noise for 60 Hz $< f_m < 1$ MHz			−80	dBc
f_0	Operating Frequency		1180		MHz
f_{REF}	Reference Frequency		62.5		kHz

Audio IF Filter (AIFF)

Table 19 lists the operating parameters of audio IF filter (AIFF) **427** in accordance with a preferred embodiment of the present invention. The audio IF signal is tapped from the output of IFAMP **422** and run through a narrow-band filter operating a 15.5 MHz with a bandwidth of 300 kHz. AIFF **427** rejects the video carrier and the burst frequency components.

TABLE 19

AIFF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_0	Center Frequency		15.5		MHz
BW	Bandwidth		300		kHz
NF	Noise Figure			10	dB

Audio IF Limiter (AIFLIM)

Table 20 lists the operating parameters of audio IF limiter (AIFLIM) **444** in accordance with a preferred embodiment of the present invention. The audio IF is limited with a 40 dB possible variation in input level based on AGC **438** and the difference between sources of the broadcast signal. The output level accuracy over this range is better than −40 dB.

TABLE 20

AIFLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	40			dB
V_{in}	Limiter Input		2		mV (p-p)

TABLE 20-continued

AIFLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
	Threshold				
V_{limit}	Limiter Output Level		0.2		V (p-p)
NF	Noise Figure			10	dB

Audio Phase Detector (APD)

Table 21 lists the operating parameters of audio phase detector (APD) **455** in accordance with a preferred embodiment of the present invention. The audio IF is compared with APS **460** output no generate a phase error signal to drive APS **460** in a delay-locked loop. APD **455** is a phase detector configured for fast loop acquisition and minimum phase error once acquired.

TABLE 21

APD Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{IN}	Voltage Input Level		0.35		V (rms)

Audio DLL Low Pass Filter (ADLPF)

Table 22 lists the operating parameters of audio delay-locked loop low-pass filter (ADLPF) **458** in accordance with a preferred embodiment of the present invention. ADLPF **458** must not respond to any frequencies above 20 Hz in order to achieve the full audio bandwidth in the detected signal. The low-pass filter accommodates this requirement.

TABLE 22

APLPF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
BW	Loop Bandwidth			20	Hz
Z_L	Closed Loop Zero		10		Hz

Audio Phase Shifter (APS)

Table 23 lists the operating parameters of audio phase shifter (APS) **460** in accordance with a preferred embodiment of the present invention. APS **460** is a voltage controlled phase shifter which is locked to the audio carrier via a delay-locked loop. APS **460** generates a 90° degree phase shift in the output signal at the aural carrier frequency of 15.5 MHz.

TABLE 23

AQO Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{OUT}	Output Voltage		0.1		V
f_0	Center Frequency		15.5		MHz
θ_S	Phase Shift		90		Deg.

Audio Quadrature Detector (AQD)

Table 24 lists the operating parameters of audio quadrature detector (AQD) **462** in accordance with a preferred

predetermined frequency and the frequency of said image signal rejected by said second mixer is less than said second operating frequency by an amount equal to said second predetermined frequency.

22. The television receiver as set forth in claim 20 wherein said first predetermined frequency is 1200 Megahertz, said second operating frequency is 1180 Megahertz, and said second predetermined frequency is 20 Megahertz.

23. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer physically located on an integrated circuit substrate having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value; and

a second mixer physically located on the same integrated circuit substrate as the first mixer and having a first input directly coupled without leaving the substrate to an output of said first mixer and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

24. The television receiver as set forth in claim 23 wherein said minimum frequency value is greater than an input cutoff frequency of an RF signal received from said RF signal source.

25. The television receiver as set forth in claim 24 wherein said first mixer subtractively mixes said RF signal and said first reference signal to thereby produce a first IF signal, wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal and said first predetermined frequency is greater than said selected carrier frequency.

26. The television receiver as see forth in claim 25 wherein said second mixer mixes said first IF signal and said second reference signal to thereby produce a second IF signal, wherein said selected carrier frequency appears in said second IF signal at a second predetermined frequency and said second predetermined frequency is less than said first predetermined frequency and said second operating frequency.

27. The television receiver as set forth in claim 26 wherein said second mixer rejects at least one image signal associated with said second operating frequency from said second IF signal.

28. A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal, wherein said RF signal mixing and said first IF signal mixing operations are accomplished on the same integrated circuit substrate.

29. The method as set forth in claim 28 including the further step of filtering out of the RF signal all frequency components above an input cutoff frequency.

30. The method as set forth in claim 29 wherein the first operating frequency is greater than the input cutoff frequency.

31. The method as set forth in claim 30 wherein the step of mixing the RF signal with the first reference signal is a subtractive mixing of the RF signal and the first reference signal.

32. The method as set forth in claim 31 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a frequency higher than the RF signal.

33. The method as set forth in claim 28 wherein the step of mixing the first IF signal with the second reference signal further includes the substep of removing from the second IF signal at lease one image signal associated with the first IF signal.

34. A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal.

35. The method as set forth in claim 34 wherein the first operating frequency is greater than the input cutoff frequency.

36. The method as set forth in claim 35 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in the RF signal appears in the first IF signal at a first predetermined frequency and wherein the first predetermined frequency is higher than the RF signal.

37. The method as set forth in claim 34 wherein the step of mixing the first IF signal with the second reference signal further includes the substep of removing from the second IF signal at least one image signal associated with the first IF signal.

38. The method as set forth in claim 34 wherein the step of mixing the first IF signal with the second reference signal is a down conversion mixing, wherein a selected carrier signal appearing in the first IF signal at a first predetermined frequency appears in the second IF signal at a second predetermined frequency and wherein the second predetermined frequency is lower than the first IF signal.

39. The method as set forth in claim 34 including the further step of filtering the second IF signal with a bandpass filter.

40. The method as set forth in claim 39 wherein the bandpass filter has a passband between 15 Megahertz and 21 Megahertz.

41. The method as set forth in claim 39 wherein the bandpass filter is a Nyquist slope filter having a cutoff frequency of 20 Megahertz.

42. A television receiver comprising:

an input filter coupled to an RF signal source;

a first amplifier coupled to an output of said input filter;

a first mixer having a first input and a second input, wherein said first input of said first mixer is coupled to an output of said first amplifier;

a first oscillator coupled to said second input of said first mixer, wherein said first oscillator generates a first reference signal having a first operating frequency;

TABLE 29

Bit Functions		
Bit Name	Function	Description
F0-15	Receiver frequency	$f = 0.0625 \cdot F + k_f$ (MHz) (F15 = MSB, F0 = LSB)
PWRDN	Circuit power-down	When high the receiver is disabled, the video and audio outputs are muted and the I^2C remains functional. The circuit draws minimum current.
AMUTE	Audio mute	Audio output is disabled when high.
VMUTE	Video mute	Video output is disabled when high.
MODE0-3	Receiver mode select	Selects receiver operating mode (unused, reserved for NTSC/PAL/SECAM switching).
VRS0-1	Video output impedance select	00 = 75Ω, 01 = 1 kΩ, 02 = Current Source.
ARS0-1	Audio output impedance select	00 = 600Ω, 01 = 10 kΩ, 02 = Current Source.
RLOCK	Receiver lock	High if AFC is locked to incoming signal
FLOCK	Synthesizer lock	High if frequency synthesizer locked
AFC0-2	AFC offset	Indicates magnitude and direction of mistuning of receiver. Three MSB's of AFC counter.
RSSI0-3	Received signal strength indicator	Indicates relative signal strength of signal. 0000 is no effective signal, 1111 is large signal. (RSSI3 = MSB, RSSI0 = LSB)
T0-7	Test data	Data used for test purposes only.
C0-15	Company code	Unique identifier for Cirrus Logic I^2C parts
R0-15	Revision code	Unique identifier for part number and revision

Bias and Control (BC)

Table 30 lists the operating parameters of bias and control logic (BC) **180** in accordance with a preferred embodiment of the present invention. Bias voltages and currents are generated to define specific levels need for each circuit element in the present invention. The control signal POWERDOWN **476** controls the power-down function via BC **180**.

TABLE 30

BC Specifications					
Parameter	Description	Min	Typ	Max	Unit
F_{osc}	Crystal Oscillator Frequency Reference		8		MHz

Although the present invention and its advantages have been described in detail, it should be understood that various changes, substitutions and alterations can be made herein without departing from the spirit and scope of the invention as defined by the appended claims.

What is claimed is:

1. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled directly to an output of said first mixer and a second input coupled to said second reference signal, wherein said second mixer is an image rejection mixer.

2. The television receiver as set forth in claim **1** wherein said first mixer is a subtractive mixer.

3. The television receiver as set forth in claim **2** wherein a minimum frequency value of said first reference signal is greater than an input cutoff frequency of an RF signal received from said RF signal source.

4. The television receiver as set forth in claim **1** wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

5. The television receiver as set forth in claim **4** wherein said first mixer subtractively mixes said first reference signal and an RF signal received from said RF signal source to thereby generate a first IF signal on said first mixer output.

6. The television receiver as set forth in claim **5** wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

7. The television receiver as set forth in claim **6** wherein said first predetermined frequency is 1200 Megahertz.

8. The television receiver as set forth in claim **6** wherein said selected carrier frequency is selected by varying said first operating frequency.

9. The television receiver as set forth in claim **1** wherein said first mixer is coupled to said receiver input by an input filter which filters out all frequency components in said RF signal above an input cutoff frequency.

10. The television receiver as set forth in claim **9** wherein said minimum frequency value of said first reference signal is greater than said input cutoff frequency.

11. The television receiver as set forth in claim **6** wherein said second mixer mixes said second reference signal and said first IF signal to thereby generate a second IF signal containing all of the channels of information as contained in said RF signal.

12. The television receiver as set forth in claim **11** wherein said second operating frequency is fixed.

13. The television receiver as set forth in claim **12** wherein said second mixer rejects from said second IF signal an image signal of said second reference signal.

14. The television receiver as set forth in claim **13** wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

15. The television receiver as set forth in claim **13** wherein said minimum frequency value of said first reference signal is greater than an input cutoff frequency of said RF signal.

16. The television receiver as set forth in claim **1** further comprising a first IF signal generated by said first mixer wherein said second mixer mixes said second reference signal and said first IF signal to generate a second IF signal containing all of the channels of information as contained in an RF signal received from said RF signal source.

17. The television receiver as set forth in claim **16** wherein said second operating frequency is fixed.

18. The television receiver as set forth in claim **17** wherein said second mixer rejects from said second IF signal an image signal of said first IF signal.

19. The television receiver as set forth in claim **18** wherein a selected carrier frequency in an RF signal received from said RE signal source appears at a first predetermined frequency in said first IF signal and at a second predetermined frequency in said second IF signal.

20. The television receiver as set forth in claim **19** wherein said second predetermined frequency is a difference in frequency between said first predetermined frequency and said second operating frequency.

21. The television receiver as set forth in claim **20** wherein said second operating frequency is less than said first

71. The television receiver as set forth in claim 70 wherein said video output buffer amplifier comprises a linear transconductance amplifier.

72. The television receiver as set forth in claim 71 wherein said video output buffer amplifier further comprises blanking means for setting an output signal of said video output buffer amplifier to a video blanking level.

73. The television receiver as set forth in claim 72 wherein said video output buffer amplifier further comprises switching means for varying an output load of said video output buffer amplifier.

74. The television receiver as set forth in claim 55 wherein said television receiver further comprises a third filter coupled to an output of said second amplifier. wherein said third filter is a narrow-band filter for filtering an audio component signal from a video output signal on said output of said second amplifier.

75. The television receiver as set forth in claim 74 wherein said third filter has a center frequency of 15.5 Megahertz and a band width of 300 Kilohertz.

76. The television receiver as set forth in claim 74 wherein said television receiver further comprises:

audio phase detector circuitry coupled to an output of said third filter; and

a third mixer having a first input and a second input. wherein said first input of said third mixer is coupled to said output of said third filter and said second input of said third mixer is coupled to an output of said audio phase detector circuitry.

77. The television receiver as set forth in claim 76 wherein said audio phase detector circuitry comprises a delay-locked loop. wherein an output signal of said delay-locked loop operates in phase quadrature with a center frequency of said audio component signal on said output of said third filter.

78. The television receiver as set forth in claim 76 wherein said television receiver further comprises a fourth filter coupled to an output of said third mixer. wherein said fourth filter is a low-pass filter with a cut-off frequency of 120 Kilohertz.

79. The television receiver as set forth in claim 76 wherein said television receiver further comprises an audio output buffer amplifier coupled to an output of said third mixer.

80. The television receiver as set forth in claim 79 wherein said audio output buffer amplifier comprises a linear transconductance amplifier.

81. The television receiver as set forth in claim 79 wherein said audio output buffer amplifier further comprises muting means for setting an output signal of said audio output buffer amplifier to zero volts.

82. The television receiver as set forth in claim 80 wherein said audio output buffer amplifier further comprises switching means for varying an output load of said audio output buffer amplifier.

83. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a first bandpass filter coupled to an output of said first mixer; the bandpass filter passing more than one channel and performing partial image rejection as well as limiting overall signal power levels to be processed by subsequent circuitry;

a second reference signal having a second operating frequency; and

an image rejection mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to said second reference signal.

84. The television receiver as set forth in claim 83 wherein said first mixer is a subtractive mixer.

85. The television receiver as set forth in claim 84 wherein a minimum frequency value of said first reference signal is greater than an input cutoff frequency of an RF signal outputted from said RF signal source.

86. The television receiver as set forth in claim 83 wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

87. The television receiver as see forth in claim 86 wherein said first mixer subtractively mixes said first reference signal and an RF signal outputted from said RF signal source to thereby generate a first IF signal on said first mixer output.

88. The television receiver as set forth in claim 87 wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

89. The television receiver as set forth in claim 88 wherein said first predetermined frequency is 1200 Megahertz.

90. The television receiver as set forth in claim 88 wherein said selected carrier frequency is selected by varying said first operating frequency.

91. The television receiver as set forth in claim 87 wherein said first mixer is coupled to said receiver input by an input filter which filters out all frequency components in said RF signal above an input cutoff frequency.

92. The television receiver as set forth in claim 91 wherein said minimum frequency value of said first reference signal is greater than said input cutoff frequency.

93. The television receiver as set forth in claim 88 wherein said second mixer mixes said second reference signal and a first filtered signal on said output of said first bandpass filter to thereby generate a second IF signal.

94. The television receiver as set forth in claim 93 wherein said second operating frequency is fixed.

95. The television receiver as set forth in claim 94 wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

96. The television receiver as set forth in claim 94 wherein said minimum frequency value of said first reference signal is greater than an input cutoff frequency of said RF signal.

97. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value;

a first bandpass filter constructed in the same substrate as said first mixer and coupled to an output of said first mixer; and

a second mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to a second reference signal having a second operating frequency. wherein said second operating frequency is fixed.

98. The television receiver as set forth in claim 97 wherein said minimum frequency value is greater than an input cutoff frequency of an RF signal outputted from said RF signal source.

99. The television receiver as set forth in claim 98 wherein said first mixer subtractively mixes said RF signal and said first reference signal to thereby produce a first IF signal, wherein a selected carrier frequency in said RF signal

a second mixer having a first input and a second input, wherein said first input of said second mixer is coupled directly to an output of said first mixer without the connection of a filter therebetween;

a second oscillator coupled to said second input of said second mixer, wherein said second oscillator generates a second reference signal having a second operating frequency; and

a second filter coupled to an output of said second mixer.

43. The television receiver as set forth in claim 42 wherein said input filter is a low-pass filter having a −3 dB cutoff frequency of 900 Megahertz.

44. The television receiver as set forth in claim 42 wherein said first amplifier is a transconductance amplifier having a noise figure between 4.5 dB and 5.4 dB.

45. The television receiver as set forth in claim 42 wherein the gain of said first amplifier is automatically controlled by an automatic gain control circuit of said television receiver.

46. The television receiver as set forth in claim 42 wherein said first mixer is an up-converter which subtractively mixes a filtered and amplified RF signal having a maximum cutoff frequency with said first reference signal to thereby generate a first IF signal, wherein a lower limit of said first operating frequency is greater than said maximum cutoff frequency.

47. The television receiver as set forth in claim 42 wherein said first oscillator is a phase-locked synthesizer and said first operating frequency has an upper limit of 2100 Megahertz and a lower limit of 1200 Megahertz.

48. The television receiver as set forth in claim 42 wherein said second mixer is an image-rejection mixer which down-converts a first IF signal outputted from said first mixer to thereby generate a second IF signal.

49. The television receiver as set forth in claim 42 wherein said second oscillator is a fixed tuned-frequency synthesizer and said second operating frequency is approximately 1180 Megahertz.

50. The television receiver as set forth in claim 42 wherein said second filter is a bandpass filter having an upper passband limit of 20.75 Megahertz and a lower passband limit of 15.8 Megahertz.

51. The television receiver as set forth in claim 42 wherein said second filter is a Nyquist slope filter.

52. The television receiver as set forth in claim 51 wherein said second filter has a −6 dB gain point at a visual carrier frequency of an IF signal generated by said second mixer.

53. The television receiver as set forth in claim 52 wherein said visual carrier frequency is approximately 20 Megahertz and said second filter has a passband corner frequency of approximately 19.25 Megahertz.

54. The television receiver as set forth in claim 52 wherein said second filter is automatically tuned by measuring the level of synchronization pulses on a input of said second filter and on an output of said second filter and adjusting a cut-off frequency of said second filter so that there is a 6 dB loss on said output of said second filter.

55. The television receiver as set forth in claim 42 wherein said television receiver further comprises:

a second amplifier coupled to an output of said second filter.

56. The television receiver as set forth in claim 55 wherein the gain of said second amplifier is automatically controlled by an automatic gain control circuit of said television receiver.

57. The television receiver as set forth in claim 55 wherein said television receiver further comprises:

a third filter coupled to an output of said second amplifier, wherein said third filter is a bandpass filter for extracting a video carrier signal from an output signal of said second amplifier; and

a third mixer having a first input and a second input, wherein said first input of said third mixer is coupled to said output of said second amplifier and said second input of said third mixer is coupled to an output of said third filter and wherein an output signal on an output of said third mixer is a video baseband signal.

58. The television receiver as set forth in claim 57 wherein said second input of said third mixer is coupled to said third filter by amplitude limiting circuitry, wherein said amplitude limiting circuitry provides a fixed-amplitude reference carrier signal to said third mixer.

59. The television receiver as set forth in claim 57 wherein said television receiver further comprises frequency detection circuitry having a first input and a second input, wherein said first input of said frequency detection circuitry is coupled to said output of said third filter and said second input of said frequency detection circuitry is coupled to a third reference signal having a third operating frequency.

60. The television receiver as set forth in claim 59 wherein said frequency detection circuitry compares the frequency of said video carrier signal to said third operating frequency to thereby produce an output control signal, wherein said output control signal is used to control said first operating frequency of said first reference signal generated by said first oscillator.

61. The television receiver as set forth in claim 57 wherein said television receiver further comprises a fourth filter coupled to said output of said third mixer and wherein said fourth filter is a low-pass filter that cuts off all signal components higher in frequency than said video baseband signal.

62. The television receiver as set forth in claim 61 wherein said television receiver further includes a clamping circuit for clamping synchronization pulses in said video baseband signal from said fourth filter.

63. The television receiver as set forth in claim 62 wherein said television receiver further comprises a noise inverter circuit coupled to said clamping circuit, wherein said noise inverter circuit removes transient noise pulses from said video baseband signal.

64. The television receiver as set forth in claim 63 wherein said transient noise pulses comprise transient noise pulses that cause said video baseband signal to drop below a blanking level of said video baseband signal.

65. The television receiver as set forth in claim 63 wherein said transient noise pulses comprise transient noise pulses that cause said video baseband signal to rise above a 108 IRE white level of said video baseband signal.

66. The television receiver as set forth in claim 57 wherein said television receiver further comprises an automatic gain control circuit coupled to said output of said third mixer, wherein said automatic gain control circuit controls the gain of said second amplifier.

67. The television receiver as set forth in claim 66 wherein said automatic gain control circuit controls the gain of said first amplifier.

68. The television receiver as set forth in claim 66 wherein said automatic gain control circuit comprises a sample-and-hold circuit coupled to a first input of a differential amplifier.

69. The television receiver as set forth in claim 68 wherein a second input of said differential amplifier is coupled to a reference voltage equal to an output blanking level of a standard NTSC video baseband signal.

70. The television receiver as set forth in claim 57 wherein said television receiver further comprises a video output buffer amplifier coupled to said output of said third mixer.

appears at a first predetermined frequency in said first IF signal and said first predetermined frequency is greater than said selected carrier frequency.

100. The television receiver as set forth in claim 99 wherein said selected carrier frequency appears at said first predetermined frequency in an output passband signal of said bandpass filter.

101. The television receiver as set forth in claim 100 wherein said second mixer mixes said output passband signal and said second reference signal to thereby produce a second IF signal, wherein said selected carrier frequency appears in said second IF signal at a second predetermined frequency and said second predetermined frequency is less than said first predetermined frequency and said second operating frequency.

102. A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a first bandpass filter to produce an output passband signal having at least four channels; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

103. The method as set forth in claim 102 including the further step of filtering out of the RF signal all frequency components above an input cutoff frequency.

104. The method as set forth in claim 103 wherein the first operating frequency is greater than the input cutoff frequency.

105. The method as set forth in claim 104 wherein the step of mixing the RF signal with the first reference signal is a subtractive mixing of the RF signal and the first reference signal.

106. The method as see forth in claim 105 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a first predetermined frequency and wherein the first predetermined frequency is higher than the selected carrier frequency.

107. The method as set forth in claim 106 wherein the first predetermined frequency appears in the output passband signal of the first bandpass filter.

108. A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a bandpass filter to produce an output passband signal having at least four channels; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

109. The method as set forth in claim 108 wherein the first operating frequency is greater than the input cutoff frequency.

110. The method as set forth in claim 109 wherein the step of mixing the amplified first filtered signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in said received RF signal appears in the first IF signal at a first predetermined frequency and wherein said first predetermined frequency is higher than the selected carrier frequency.

111. The method as set forth in claim 110 wherein the selected carrier signal appears in the output passband signal at the first predetermined frequency.

112. The method as set forth in claim 108 wherein the step of mixing the output passband signal with the second reference signal is a down conversion mixing, wherein a selected carrier signal appearing at a first predetermined frequency in the output passband signal appears in the second IF signal at a second predetermined frequency and wherein said second predetermined frequency is lower than said first predetermined frequency.

113. The method as set forth in claim 108 including the further step of filtering the second IF signal with a second bandpass filter.

114. The method as set forth in claim 113 wherein the second bandpass filter has a passband between 15 Megahertz and 21 Megahertz.

115. The method as set forth in claim 114 wherein the second bandpass filter is a Nyquist slope filter having a cutoff frequency of 20 Megahertz.

116. The method of extracting a single communications channel from a large number of such channels wherein each such channel has a defined frequency range, and wherein the large number of channels are communicated within an upper and a lower frequency boundary, the method comprising the steps of:

from an input signal containing all of the large number of channels removing all frequencies greater than the upper frequency boundary;

translating all of the channels contained within the input signal to a translated frequency range such that a desired single channel to be extracted from the input signal is positioned centrally within the translated frequency range, the translated frequency range being positioned at frequencies higher than the upper frequency boundary;

removing from the translated input signal all channels extending from the center of the translated frequency range four channels or more in each frequency direction;

translating the remaining channels within the input signal to a second translated frequency range, such that the desired single channel to be extracted from the input signal is positioned centrally within the second translated frequency range having a low end frequency boundary defined as high enough not to cause phase distortion in any subsequent filtering and a high end frequency boundary defined as a frequency beyond which a filter with proper bandwidth, noise and distortion can be integrated in a substrate; and

removal of all remaining frequencies which are not part of the desired single channel.

117. The method set forth in claim 116 wherein the last-mentioned removal step includes a receiver attenuation characteristic such that the relative amplitude of low and high frequencies within the desired single channel are corrected for vestigial sideband demodulation.

118. The method set forth in claim 116 wherein the low end and high end boundaries of the second translated frequency range are defined by the implementation of an integrated circuit continuous time filter capable of achieving single channel TV signal extraction and receiver attenuation characteristics.

119. The television receiver as set forth in claim 42 wherein said input filter limits the frequencies passed to those at or below the highest channel in a desired frequency band.

120. A television receiver comprising:

a receiver input for receiving an RF signal;

an input filter coupled to said receiver input and operating to remove all frequency components in said RF signal above an input cutoff frequency to produce a filtered RF signal;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to an output of said input filter and a second input coupled to said first reference signal;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled to an output of said first mixer and a second input coupled to said second reference signal, wherein said second mixer is an image rejection mixer.

121. The television receiver of claim 120 wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

122. The television receiver of claim 121 wherein said minimum frequency value is greater than said input cutoff frequency.

123. The television receiver of claim 120 wherein said first mixer subtractively mixes said first reference signal and said filtered RF signal to thereby generate a first IF signal at said first mixer output.

124. The television receiver of claim 120 wherein said second mixer mixes a first IF signal from said first mixer and said second reference signal to generate a second IF signal and wherein said second mixer rejects an image signal of said second reference signal from said second IF signal.

125. The television receiver of claim 120 wherein said second mixer mixes a first IF signal from said first mixer and said second reference signal to generate a second IF signal and wherein said second mixer rejects an image signal of said first IF signal from said second IF signal.

126. The television receiver of claim 120 wherein said first mixer generates a first IF signal and wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

127. The television receiver of claim 126 wherein said second mixer generates a second IF signal and wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

128. The television receiver of claim 120 wherein said first mixer and said second mixer are physically located on the same integrated circuit substrate.

* * * * *

TAB 2

PATENT APPLICATION SERIAL NO. 08/426080

U.S. DEPARTMENT OF COMMERCE
PATENT AND TRADEMARK OFFICE
FEE RECORD SHEET

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TUNE 048156

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U.S. PATENT APPLICATION

SERIAL NUMBER	FILING DATE	CLASS	GROUP ART UNIT
08/426,080	04/21/95	348	2602

APPLICANT

ROBERT R. ROTZOLL, ALLEN, TX.

CONTINUING DATA********************
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STATE OR COUNTRY	SHEETS DRAWING	TOTAL CLAIMS	INDEPENDENT CLAIMS	FILING FEE RECEIVED	ATTORNEY DOCKET NO.
TX	4	116	9	$3,298.00	2836-P223US

ADDRESS

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DALLAS TX 75270

TITLE

MONOLITHIC TELEVISION TUNER

This is to certify that annexed hereto is a true copy from the records of the United States Patent and Trademark Office of the application which is identified above.

By authority of the
COMMISSIONER OF PATENTS AND TRADEMARKS

Date

Certifying Officer

TUNE 048157

[redacted]
426080

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24 APR 1995
TRADEMARK

APPLICATION FOR U.S. PATENT
TRANSMITTAL FORM

THE COMMISSIONER OF PATENTS
AND TRADEMARKS
Washington, D.C. 20231

Docket No. 2836-P223US

Sir:

Transmitted herewith for filing is the patent application of:

Inventor(s): Robert Rudolf Rotzoll

For: MONOLITHIC TELEVISION TUNER

Enclosed are: 4 Sheets of Informal Drawings

An assignment of the invention to:
Cirrus Logic, Inc.

FEE CALCULATION					FEE
	Number		Number Extra	Rate	Basic Fee $ 730.00
Total Claims	116	- 20 =	96	x $22 =	$ 2,112.00
Independent Claims:	9	- 3 =	6	x $76 =	$ 456.00
Multiple Dependent Claims:					$
				Total Filing Fee =	$ 3,298.00

Enclosed is a check in the amount of $3,298.00. Please charge any additional fees or credit any overpayment to Deposit Account No. 23-2426 of Winstead Sechrest & Minick P.C. A duplicate copy of this sheet is enclosed.

Please return the original Assignment document to the undersigned attorney for Applicant following recordal of same and address it to: WINSTEAD SECHREST & MINICK P.C., 5400 Renaissance Tower, 1201 Elm Street, Dallas, Texas 75270.

April 21, 1995
Date

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Attorney for Applicant
Registration No. 24,745

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233:2836-P223US

08/426080

(2836-P223US) PATENT

MONOLITHIC TELEVISION TUNER

ABSTRACT OF THE DISCLOSURE

There is disclosed a fully integrated television receiver for receiving a standard antenna or cable input and outputting a standard video baseband signal and a standard audio baseband signal. The receiver employs an up-conversion mixer and a down-conversion image-rejection mixer in series to produce an on-chip IF signal. Audio detection of the IF signal is performed via an audio delay-locked loop operating in quadrature to the audio FM signal. Video detection of the IF signal is performed by frequency mixing the IF signal with the extracted video carrier of the IF signal. Variable load drive capability is provided for both the video output and the audio output.

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10/04/96

ATTORNEY DOCKET NO. 4.0c1-P023US-954919

PATENT APPLICATION

IN THE U.S. PATENT AND TRADEMARK OFFICE

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GROUP 260

Inventor(s):	Robert Rudolf Rotzoll		
Serial No.:	08/426-080	Examiner:	Sherrie Hsia
Filing Date:	April 21, 1995	Group Art Unit:	2602
Title:	MONOLITHIC TELEVISION TUNER		

ASSISTANT COMMISSIONER
FOR PATENTS
Washington, D.C. 20231

Sir:

TRANSMITTAL LETTER FOR AMENDMENT

Transmitted herewith is/are the following in the above-identified application:

(X) Response/Amendment ( ) Petition to extend time to respond

( ) New fee as calculated below ( ) Supplemental Declaration

(X) No additional fee (Address envelope to "Box NON-FEE Amendment")

( ) Other: ______

CLAIMS AS AMENDED BY OTHER THAN A SMALL ENTITY						
(1) FOR	(2) CLAIMS REMAINING AFTER AMENDMENT	(3) NUMBER EXTRA	(4) HIGHEST NUMBER PREVIOUSLY PAID FOR	(5) PRESENT EXTRA	(6) RATE	(7) ADDITIONAL FEES
TOTAL CLAIMS	37	MINUS	37	= 0	X $22	$ 00.00
INDEP. CLAIMS	4	MINUS	4	= 0	X $80	$ 00.00
[ ] FIRST PRESENTATION OF A MULTIPLE DEPENDENT CLAIM					+$250	$
EXTENSION FEE	1ST MONTH $110	2ND MONTH $390	3RD MONTH $930	4TH MONTH $1470		$
					TOTAL ADDITIONAL FEE FOR THIS AMENDMENT	$ 00.00

Enclosed is a check in the amount of $00.00. At any time during the pendency of this application, please charge any fees required or credit any overpayment to Deposit Account 06-2380 pursuant to 37 CFR 1.25. Additionally, please charge any fees to Deposit Account 06-2380 under 37 CFR 1.19, 1.20 and 1.21. A duplicate copy of this sheet is enclosed.

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Date of Deposit: 10-1-96

Typed Name: LINDA CUCCIA

Signature: Linda Cuccia

Respectfully submitted,

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0335604

(2836-P223US) PATENT

MONOLITHIC TELEVISION TUNER

TECHNICAL FIELD OF THE INVENTION

This invention relates to television tuner circuits and more particularly to a highly integrated television tuner fabricated in a single microcircuit device.

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PATENT

BACKGROUND OF THE INVENTION

One of the most significant costs in television manufacturing is the cost of the tuner. The typical cost of a television (TV) tuner is in the neighborhood of $15.00, which, relative to the cost of the entire television set, is very substantial. Part of the solution to reducing tuner cost is to reduce the number of components in the tuner.

Traditionally, tuners have been comprised of two basic components. The first component performs high frequency to intermediate frequency (RF to IF) conversion. Subsequently, the second component performs IF to baseband conversion. The TV tuner was originally designed for broadcast television reception within a television set, which is essentially a stand-alone unit containing a cathode ray picture tube. So, TV tuners were originally integral parts embedded in a single-purpose device.

Presently, however, state-of-the-art consumer electronic devices use TV tuners that are not a built-in part of a television set. The tuner is a separate element that is connected to a cathode ray picture tube at some point, but the tuner is not an integral part of the monitor. For example, TV tuners may be fabricated on circuit boards and then installed in personal computer (PC) systems, thereby allowing the PC to function as a television set. These tuners convert a radio frequency television signal into a baseband (or low frequency) video signal which can then be passed on to other elements in the PC for video processing applications.

The circuit component that performs the RF-to-IF conversion typically comprises one or two integrated

4

circuits and numerous discrete elements -- inductors, capacitors and/or transistors. The IF-to-baseband conversion typically includes another integrated circuit, several filter elements, such as ceramic filters and SAW filters, a series of tuning and control elements, such as resistors and potentiometers, variable inductors and/or capacitors, and some other additional external components. Thus, the complexity of the tuner is fairly high and typically there may be between 100 and 200 elements on a circuit board. Furthermore, state-of-the-art TV tuners still require that each tuner be aligned by manual tuning before leaving the factory. This manual tuning is one of the most expensive costs associated with the manufacturing process and an important factor in the cost of tuners.

Broadcast television tuners of the past have gone through an evolution over a period of more than 60 years. The earliest tuners utilized vacuum tube technology and required that the minimum number of vacuum tubes possible be used due to their cost, power consumption and dimensions. Therefore, passive components, such as resistors, capacitors, inductors and transformers, were used as much as possible in most designs. This style of design continued until about 1960 when TV tuner components, particularly vacuum tubes, began to be replaced by bipolar and MOS transistors. However, the active device count still defined the cost and size limits of TV tuners and active device count minimization continued.

In the early 1970's the integrated circuit became viable as an element in the television tuner and the design techniques were dramatically changed. Many functions of the tuner utilizing only one tube or transistor were being replaced with 4 to 20 individual

transistors which could perform the same function with better precision, less space, less power, less heat generation and lower cost. The introduction of the integrated circuit was gradual, first encompassing only low frequency elements and then eventually high frequency active elements. Nonetheless, many passive elements external to the integrated circuits remained in TV tuner designs.

One advance, the SAW (surface acoustic wave) filter, made a significant change in that several manually tuned inductors and capacitors could be removed from the tuners and receive-filtering performance could be improved within a much smaller space and at reduced cost. However, the SAW filter, which is fabricated on a ceramic substrate, cannot be integrated on a silicon wafer with the rest of the active circuitry and must therefore remain a discrete component in the final design. The trend of the 1980's was to miniaturize all of the passive components and simplify their associated manual tuning at the factory. In recent years, TV tuners have been reduced in size from requiring fairly large enclosures, about 2" x 5" x 1", to much smaller enclosures, about 1/2" x 2" x 3/8". There is a high premium placed on small size because TV tuners are being used in smaller and smaller computers, television sets and VCRs. As the equipment in which tuners are used becomes smaller, the size of the TV tuner must decrease also.

As the size of the tuner goes down, and as tuners are used in a wider variety of devices, cost becomes more critical and must be reduced as much as possible in order not to represent a large portion of the final product cost. When a tuner is used in a television set, the tuner size is less critical because the

television set inherently has a large mass. But when a tuner is used in other electronic equipment, space becomes a premium and the footprint of the tuner becomes critical.

Accordingly, it is one object of the invention to provide a TV tuner which has a relatively low cost and a small footprint for use on a printed circuit board.

It is another object of the present invention to provide a TV tuner that meets or exceeds the performance of state-of-the-art TV tuners while at the same time reducing the number of external components needed, thereby decreasing the complexity of the printed circuit board and the amount of circuit board area needed by the TV tuner.

It is the further object of the present invention to allow for computer control of the TV tuner by a serial bus so that the TV tuner may be controlled by a microcontroller imbedded in the television set, personal computer, or other video device.

It is the further object of the present invention to provide a TV tuner with computer-controlled output impedance characteristics to accommodate different load specifications.

SUMMARY OF THE INVENTION

These and other problems have been solved by integrating a TV tuner that takes a broad band of frequencies as an input and performs video demodulation over a very narrow band width. To accomplish this, an architecture was chosen to perform an up-conversion of the RF input signal to a higher internal frequency, which allows the present invention to have minimal filtering on the input stages of the receiver. The present invention is therefore able to operate without variable tuned filtering on its input, with the exception of a single fixed tuned filter. This eliminates the need for precisely controlled variable tuned filters which must be mechanically aligned during manufacture and are subject to variation in performance due to age, temperature, humidity, vibration and power supply performance. This was a critical drawback of previous tuners that had to be eliminated because it is a source of tremendous error and distortion, as well as complexity.

Furthermore, it is advantageous to have the up-conversion performed on-chip, because eliminating the inputting or outputting of high frequency signals to or from the integrated circuit of the present invention will avoid drive capability problems associated with high frequency signals and noise coupling problems associated with integrated circuit external interconnections.

The present invention next performs a down-conversion from the higher frequency using an image rejection mixing scheme that provides a tightly controlled down-conversion with very little added distortion. This also further minimizes the on-chip

filtering effects by minimizing the amount of filtering that must be done at high frequency. The result is that high frequency operations within the present invention are limited to the initial stage inside the chip and are performed at a very low impedance level in order to minimize distortion. After the down conversion, all remaining filtering is performed in continuous time mode filters that are realizable in an integrated circuit.

Another technical advantage of the present invention is that no manual tuning of the final design is required. There are no external components needed to perform adjustments, as there is in the prior art.

A further technical advantage of the present invention is that the output impedance is selectable, based on system requirements, thereby minimizing power dissipation for certain applications. The present invention can drive a coaxial cable with a low output impedance driver or, for board level applications, the present invention can be configured to drive a high output impedance. The present invention can also match a very precise load impedance defined by a single external element.

The present invention advantageously utilizes much less board space than previous designs (on the order of 5% to 10% of the prior art designs) and dissipates only about one-fifth of the power. The present invention also advantageously operates on a single voltage level, as opposed to two or three levels for previous designs.

A further technical advantage of the present invention is that the need for a metal enclosure is reduced. Integration, by itself, allows for sufficient shielding to meet interference standards.

The monolithic television (MTV) tuner embodied in the present invention is intended to replace the TV tuner modules presently used in most broadcast television receiver devices. The MTV tuner integrates all of the functional elements of a tuner, except for the quartz crystal frequency reference and power supply bypass capacitors. The level of integration of the present invention dramatically reduces the cost of the basic TV tuner and enhances its manufacturability and reliability.

The TV tuner of the present invention is controlled externally by a computer or controller via a digital serial bus interface (I^2C). A preferred embodiment of the present invention provides an antenna input capable of being connected directly to a standard coaxial cable, thereby allowing both antenna and cable television applications. Furthermore, the baseband video and audio outputs are designed for high impedance or low impedance applications. The high impedance mode is a driver for short interconnections on a printed circuit and gives the benefit of low operating power. The low impedance mode drives an industry standard studio cable interface requiring greater supply power.

A preferred embodiment of the present invention is designed to operate on frequencies used for both over-the-air broadcasts and cable television with National Television Standards Committee (NTSC) encoded video. Receiver sensitivity is set to be limited by the antenna noise temperature for VHF systems. The present invention also employs a wide-range automatic gain control (AGC).

The baseband video output of the present invention is leveled, or has minimal variation in video amplitude with respect to antenna RF signal level, and can be

configured via software for either low impedance studio standard loads or for high impedance loads with lower operating current. Audio output is broadband composite to allow connection to an external MTS decoder. The audio output can be configured for low impedance studio standard loads or high impedance loads via software control.

Control is accomplished via an I^2C bus interface. The bias and control circuits in a preferred embodiment of the present invention contain internal registers which can be updated via the control bus for operating frequency, video and audio transmission standards such as NTSC, PAL, SECAM and MTS, power, and test modes. Status of the bias and control circuits can be checked via a status register accessible through the I^2C bus interface. Status data include AFC error, channel lock and received signal strength indicator.

The operating frequency of the present invention is referenced to an external crystal or reference frequency generator. A minimum of external components are used with the present invention and no tuning of any components is required.

The present invention may be implemented in Bipolar, BiCMOS, or CMOS processes, however, a preferred embodiment of the present invention employs a BiCMOS process to reduce the difficulty in developing the design by allowing maximum flexibility.

A preferred embodiment of the present invention performs sound processing after IF amplification and prior to video demodulation. This method is known as "quasi-split" sound and is not used in the majority of television receivers for cost reasons. Quasi-split sound does not have a "buzz" in the audio signal during periods when the video picture is very white. In the

integrated circuit of the present invention, quasi-split sound is a negligibly small cost and thus may be integrated in all embodiments of the present invention. Integration of quasi-split sound further reduces component count and does not require external manual adjustments.

The foregoing has outlined rather broadly the features and technical advantages of the present invention in order that the detailed description of the integrated television tuner that follows may be better understood. Additional features and advantages of the monolithic television tuner will be described hereinafter which form the subject of the claims of the invention. It should be appreciated by those skilled in the art that the conception and the specific embodiment disclosed may be readily utilized as a basis for modifying or designing other structures for carrying out the same purposes of the present invention. It should also be realized by those skilled in the art that such equivalent constructions do not depart from the spirit and scope of the invention as set forth in the appended claims.

BRIEF DESCRIPTION OF THE DRAWINGS

For a more complete understanding of the present invention, and the advantages thereof, reference is now made to the following descriptions taken in conjunction with the accompanying drawings, in which:

Sub B1

FIGURE 1 is a pin layout diagram of an integrated circuit in accordance with the present invention;

FIGURE 2 is a high level block diagram of the functional components of the present invention;

Ins a1

a1 ~~FIGURE 3 is a detailed block diagram of a~~ state-of-the-~~art~~ television tuner found in the prior art; and

FIGURE 4 is a detailed block diagram of the ~~circuit components of the present invention~~.

Ins a2

DETAILED DESCRIPTION OF THE PRIOR ART

Before discussing the monolithic television tuner of the present invention, it will be useful to discuss a state-of-the-art television tuner found in the prior art.

While there have been theoretical proposals to integrate TV tuners in a single microcircuit, none are known to have been implemented. The next best definition of the known prior art, then, is a highly miniaturized, but not fully integrated, tuner as shown in FIGURE 3.

FIGURE 3 depicts a functional electrical block diagram of a present state-of-the-art TV tuner configuration. Television tuner 300 is constructed in a single metallically shielded assembly containing a printed circuit board on which all of the associated tuner components are mounted and electrically connected. TV tuner 300 is designed to be a module mounted on other printed circuit boards to allow for direct connection of the input and output signals to their appropriate terminations within the television receiving system. The metal shield is used to keep undesired external signals from interfering with the operation of the TV tuner 300 and to prevent TV tuner 300 from radiating signals that interfere with the operation of external devices.

Prior art television tuner 300 is comprised of three integrated circuits: preamplifier and mixer 305, IF and baseband signal processor 310 and frequency synthesizer and Inter Integrated Circuit (IIC or I^2C) bus interface 315 . Television tuner 300 is also comprised of a plurality of discrete components, including bandpass and image reject notch filter 304,

bandpass and image reject notch filter 312, surface acoustic wave (SAW) filter 316, video carrier filter 324, and audio carrier phase shifter 360.

Television tuner 300 receives a standard television RF signal from either antenna 302 or a cable system connection (not shown) through bandpass and image reject notch filter 304. Bandpass and image reject notch filter 304 limits the signals entering TV tuner 300 so that a minimum number of undesired signals exist in TV tuner 300. Filter 304 therefore limits the image response caused by the first mixer, described later. Filter 304 also attenuates signals not in a fairly narrow (100 MHz) range about the desired signal. Finally, known interference signals, such as FM broadcast, shortwave service signals, signals in the intermediate frequency band and Citizen Band radio signals, are specifically rejected by filter 304.

Preamplifier 306 of preamplifier and mixer 305 receives the output of bandpass and image reject notch filter 304 and raises the signal level (10 dB) with minimum increase in the noise level (typically 8 - 10 dB). The gain of preamplifier 306 is controlled by automatic gain control (AGC) 338, so that when a very strong signal enters TV tuner 300, overall gain is reduced, resulting in less distortion in the preamplifier than without the gain reduction.

The output of preamplifier 306 is sent to bandpass and image reject notch filter 312, with the same basic requirement of minimizing the passage of potential interference signals. Filter 312 is external to preamplifier and mixer 305 and is comprised of a plurality of discrete elements, including capacitors, inductors and varactor diodes.

The output of bandpass and image reject notch filter 312 is then sent back to mixer 308 in preamplifier and mixer 305. Mixer 308 mixes the output of filter 312 with the output of a local oscillator, frequency synthesizer 342, which has a frequency chosen to be higher than the desired receiver carrier by 45.75 MHz. Thus, the output of mixer 308 is 45.75 MHz. There also is an image signal due to mixer 308 at 91.5 MHz above the input frequency, which is removed by filter 304 and filter 312. Therefore, as the frequency of frequency synthesizer 342 is tuned to receive signals of different carrier frequencies, the bandpass and image reject filters 304 and 312 must also be tuned to properly pass only the desired signals and not the mixer images.

Frequency synthesizer 342 receives an input frequency reference signal (usually 16 bits) and outputs the status signals AUTOMATIC FREQUENCY CONTROL (AFC) ERROR and FREQUENCY (FREQ) LOCK. Additionally, a tuning signal which is used by the voltage controlled oscillator (VCO) in frequency synthesizer 342 is output from frequency synthesizer 342 to bandpass and image reject notch filters 304 and 312. A local oscillator signal is output from frequency synthesizer 342 to mixer 308.

The 45.75 MHz output signal of mixer 308 then passes through SAW (surface acoustic wave) filter 316, which limits the bandwidth of the signal to only one (1) channel (6 MHz for NTSC standard) and applies a linear attenuation in frequency known as the Nyquist slope around the visual carrier frequency. The linear attenuation by SAW filter 316 converts the signal from a vestigial sideband signal to one which is equivalent to a single sideband with a carrier, so that the

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frequency response of the signal after demodulation is flat over the video bandwidth. SAW filter 316 is very "lossy" (on the order of 25 dB), so the input to SAW filter 316 is amplified by a preamplifier (not shown) by a corresponding amount to minimize noise effects.

The output of SAW filter 316 is input to intermediate frequency (IF) amplifier 320 in IF and baseband signal processor 310. IF amplifier 320 provides most of the overall gain of TV tuner 300 and receives gain control from AGC 338.

The output of IF amplifier 320 is sent to video detector 322 and is also sent off-chip to external video carrier filter 324. This is the stage at which video demodulation is performed. Video detector 322 is essentially a mixer with the local oscillator input connected to the output of video carrier filter 324 through carrier amplitude limiter 326. The output of the carrier limiter 326 is an in-phase representation of the video carrier signal without any modulation applied to it. The output of carrier limiter 326 is received by video detector 322, which mixes the output of carrier limiter 326 with the output of IF amplifier 320.

AFC frequency discriminator 340 is used in the prior art device to detect the difference between the carrier frequency contained in the output of carrier limiter 326 and a known valid carrier frequency reference. The output signal on the output of AFC frequency discriminator 340 is an error signal which is used to drive frequency synthesizer 342 in a direction that will reduce the error between the output of carrier limiter 326 and the known valid carrier frequency reference.

The output of the video detector 322 is a baseband video signal combined with several high frequency mixing artifacts. These artifacts are removed by a video baseband filter 330. The output of video baseband filter 330 is fed to synchronization pulse clamp (sync clamp) 332, which sets the level of the sync pulses to a standard level.

Next, the output of sync clamp 332 is sent to noise inverter 334, which removes large noise spikes from the signal. The output of noise inverter 334 is sent to video buffer 336, which is configured to drive fairly high circuit board impedances of approximately 1000 to 2000 ohms.

The output of noise inverter 334 is also sent to AGC (automatic gain control) 338, which compares the level of the synchronization pulses to the signal blanking level to measure the incoming signal strength and generates a gain control signal which is used by IF amplifier 320 and RF preamplifier 306 to dynamically adjust the gain of the TV tuner 300 for the correct level at the final output.

The audio signal is an FM signal which follows the same path as the video through video detector 322. At the output of video detector 322, the audio signal appears as a subcarrier at 4.5 MHz, due to the fact that the audio signal comes into prior art TV tuner 300 4.5 MHz higher in frequency than the desired video carrier. The audio subcarrier is passed on to an FM quadrature demodulator. The FM quadrature demodulator is comprised of a mixer, audio second detector 350, and a 90 degree (at 4.5 MHz) phase shifter, audio carrier phase shifter 360. The output of the audio second detector 350 is a baseband audio signal, which is filtered by lowpass (30kHz) filter 352 to remove any

undesired high frequency components. The output of lowpass filter 352 is finally passed on to audio buffer 354, which drives an audio amplifier that ultimately drives a speaker. Serial digital interface 344 receives SERIAL DATA and SERIAL CLOCK inputs to provide control and update status for the prior art television receiver.

Baseband and image reject notch filters 304 and 312 are typically comprised of a plurality of capacitors, inductors and varactor diodes. Video carrier filter 324 is usually comprised of three discrete elements: an inductor and two capacitors. Likewise, audio carrier phase shifter 360 is also comprised of an inductor and two capacitors. In addition to the circuit elements shown as discrete components outside of circuit elements 305, 310 and 315 in FIGURE 3, other discrete components (not shown) are connected to IF and baseband signal processor 310 and frequency synthesizer 342 for tuning purposes. Frequency synthesizer 342 is typically tuned by several external capacitors, inductors and/or varactor diodes. Video buffer 336 and audio buffer 354 will also typically employ external discrete elements, such as resistors, capacitors and/or transistors. Video baseband filter 330 and low pass filter 352 may also employ external inductors and capacitors.

DETAILED DESCRIPTION OF THE INVENTION

FIGURE 1 depicts a high-level block diagram of the functional elements contained in a preferred embodiment of TV tuner 100. The RF signal received from the antenna or other source is converted by RF converter 110 to an intermediate frequency (IF) signal and sent to IF processor 120. The output of IF processor 120 is sent to Vestigial Sideband (VSB) detector 130 and video processor and driver 150 to produce the video output signal of the present invention. The video signal is detected via a synchronous AM demodulator.

The output of IF processor 120 is also sent to FM detector 140 and audio processor and driver 160 in order to produce the audio output signal of the present invention. Audio detection is performed by a quadrature FM detector utilizing a phase-locked loop for the quadrature reference. I^2C interface 170 receives I^2C DATA and I^2C CLOCK signals from the I^2C interface bus. Frequency reference 190 is an oscillator that is synchronized to the external reference crystal. Bias and control logic circuit 180 defines internal bias voltages and currents and maintains the status and control registers of the television tuner of the present invention.

FIGURE 2 illustrates a pin layout diagram of a preferred embodiment of an integrated circuit in accordance with the present invention. Integrated circuit 200 is connected to a supply voltage on pins 11 and 20 (VCC1 and VCC2) and is grounded on pins 3, 4, 6-10, 14 and 17 (GND1-GND9). An input reference signal from an external crystal oscillator is supplied on pins 1 and 2 (CRYSTAL 1 and CRYSTAL 2/REF IN). Integrated b circuit 200 ~~100~~ is connected to an IIC (I^2C) interface bus

on pin 16 (SERIAL CLOCK) and pin 15 (SERIAL DATA). The RF input from an antenna or other source is connected to pin 5 (RF INPUT). Integrated circuit 200 outputs video and audio signals on pins 18 and 19 (VIDEO and VIDEO REF) and pins 12 and 13 (AUDIO and AUDIO REF).

FIGURE 4 is a detailed electrical block diagram of TV tuner 100 in accordance with a preferred embodiment of the present invention. FIGURE 4 depicts an up-converting dual conversion superheterodyne receiver. The Vestigial Sideband (VSB) encoded video signal is processed via a Nyquist slope receiver attenuation characteristic filter prior to detection.

The RF signal enters TV tuner 100 from the antenna 402 (or cable, not shown) and is passed through RF low-pass filter (RFLPF) 404 to limit the incoming band to below 900 MHz. The filtered RF signal is amplified up to 20 dB by a gain-controlled low-noise transconductance amplifier (LNTA) 406.

The input signal received by antenna 402, filtered by RFLPF 404 and amplified by LNTA 406 is the standard broadcast television spectrum. Channels 2-13 in the VHF band extend from 54 Megahertz to 216 MHz and channels 14-83 in the UHF band extend from 470 MHz to 890 MHz. Each channel in the VHF and UHF band has a 6 MHZ bandwidth with the video carrier frequency located 1.25 MHz above the lower band edge, a color carrier frequency located 3.58 MHz above the video carrier, and an audio carrier frequency located 4.5 MHz above the video carrier. For example, Channel 2 has a 6 MHz bandwidth from 54 to 60 MHz, a video carrier at 55.25 MHz, a color subcarrier at 58.83 MHz and an audio carrier at 59.75 MHz.

The output of first local oscillator (LO1) 450, operating between 1200 and 2100 MHz, is mixed in first

mixer (MIX1) 408 with the RF signal to generate a first IF video carrier frequency of 1200 MHz. This approach leads to minimum distortion due to mixer images and harmonic mixing. The first IF is crudely filtered by the bandwidth limitation of first mixer 408 to minimize harmonic effects.

In a basic mixer scheme, a mixer receives two inputs, an RF input and a local oscillator (LO) input, and generates an IF output. The RF input has the general form:

[1] $V_{RF} = A \cos(\omega_{RF} t)$,

the local oscillator input has the general form:

[2] $V_{LO} = B \cos(\omega_{LO} t)$,

and the IF output is given by:

[3] $V_{IF} = V_{RF} \bullet V_{LO}$

$= AB/2[\cos(\omega_{RF} - \omega_{LO})t + \cos(\omega_{RF} + \omega_{LO})t]$.

In the frequency domain, the IF frequencies are given by: $f_{IF} = |f_{RF} \pm f_{LO}|$. In realizable mixers, there are various non-linearities which cause the generation of harmonics at multiples of the RF and IF frequencies. Thus, the IF spectrum contains harmonics at $f_{IF} = mf_{RF} \pm nf_{LO}$, where m and n are integers.

For example, if the output of local oscillator 450 is 400 MHz and the desired input signal to be demodulated from the antenna is 800 MHz, a basic mixer would perform an additive mix to obtain a 1200 MHz output at the output of mixer 408. However, since frequency synthesizer 450 also has a harmonic at 800 MHz, and a signal exists on the antenna at 400 MHz, these signals are also added to produce an 800 MHz output that would interfere with the desired signal to be demodulated at 800 MHz.

Mixer 408 is a subtractive mixer that will subtract the frequency of local oscillator 450 from the

input RF spectrum between 0 Hz to 900 MHz that is filtered by RFLPF 404. For example, if the desired frequency to be demodulated in the RF input is 400 MHz, the local oscillator would be set to 1600 MHz and a basic mixer would generate two IF outputs:

[4] $f_{IF} = |400 - 1600| = 1200$ MHz and

[5] $f_{IF} = |400 + 1600| = 2000$ MHz.

Since mixer 408 is a subtractive mixer, the 2000 MHz signal will be filtered out and only the 1200 MHz signal will be passed on to mixer 410. Also, since mixer 408 is a subtractive mixer, the incoming spectrum is inverted on the output, so that the video carrier is 1.25 MHz below the upper edge of the 6 MHz channel bandwidth, the color carrier is 3.58 MHz below the video carrier and the aural carrier is 4.5 MHz below the video carrier. For example, the 6 MHz bandwidth of channel 2 from 54-60 MHz would appear inverted between 1201.25 MHz and 1195.25 MHz. The channel 2 video carrier would appear at 1200 MHz, the color carrier would appear at 1196.42 and the aural carrier would appear at 1195.5 MHz.

As noted above, it is a characteristic of both the input RF signal received by mixer 408 from LNTA 406 and of frequency synthesizer 405 that harmonic signals exist at multiples of the desired frequencies. By using a subtractive mixer and a sufficiently high-frequency output from oscillator 450, the harmonics on the output of frequency synthesizer 450 will all be more than 1200 MHz higher in frequency than any of the input frequencies existing in the 0 Hz to 900 MHz range from antenna 402.

For example, the lowest output of local oscillator 450 is 1200 MHz. The first harmonic of this signal exists at 2400 MHz. Since the highest output signal

passing through LNTA 406 is 900 MHz, the subtractive mix of the 2400 MHz harmonic and the 900 MHz signal from antenna 402 would yield a signal of |900 - 1200 | = 1300 MHz. This is higher than the desired 1200 MHz output from mixer 408, and may therefore be filtered out by low-pass filtering. Using a local oscillator whose output is higher that the RF input frequency is known as high-side injection.

The first IF signal of 1200 MHz is mixed in second mixer (MIX2) 410, which is an image-rejection mixer, with the fixed 1180 MHz reference output of second local oscillator (LO2) 412 to generate the second IF at 20 MHz visual carrier. Because the RF input signal is lower in frequency than the LO reference, the mixing of the two signals will result in a down conversion of the RF input. The IF frequencies, given by $f_{IF}= |f_{RF} \pm f_{LO}|$, have carriers at f_{IF}= 1200-1180 =20 MHz and at f_{IF}= 1200+1180 =2380 MHz. The image-rejection mixer specifically rejects signal energy in the first IF signal that is in the area of 1160 MHz, which would also yield an undesired signal at $f_{IF} = |1160- 1180 |=$ 20 MHz.

The image rejection mixer receives an input signal from MIX1 that has the general form:

[6] $V_{RF} = A \cos(\omega_{RF}t)$,

and the local oscillator signal from LO2 is split into two phase quadrature signals having the general form:

[7] $V_{LOI} = B \cos(\omega_{LO}t)$ and

[8] $V_{LOQ} = B \sin(\omega_{LO}t)$.

The desired output signal, V_{IF}, resulting from the mixing of the RF input and the phase quadrature signals has the general form:

[9] $V_{IF} = AB \cos(\omega_{RF}-\omega_{LO})t$.

The desired mixer signal at $\omega_{IF} = \omega_{RF} - \omega_{LO}$ appears in the mixer output, while the undesired mixer image signal at $\omega_{IF} = \omega_{RF} + \omega_{LO}$ (-1160 + 1180 = 20) is rejected.

FIGURE 5 depicts an alternate embodiment of the RF front-end (i.e., RF converter 110 and IF processor 120) of the present invention. First IF filter (FIFF) 509 between mixer 408 and mixer 510 filters out all frequency components of the image falling on 1160 MHz, such that mixer 510 is a basic mixer similar to mixer 408, rather than an image rejection mixer. While the image rejection mixer is easier to integrate, it has accuracy limitations, and noise and distortion figures may suffer.

Returning to FIGURE 4, the second IF signal is next filtered for the final video bandwidth by second IF filter (SIFF) 420, a low-distortion continuous-time bandpass filter. SIFF 420 also comprises an automatically tuned low-pass Nyquist slope filter (NSF) for VSB balancing. The -6 dB point on the slope of the NSF is maintained at the visual carrier frequency (20MHz).

The output of SIFF 420 is coupled to IF amplifier (IFAMP) 422, a gain-controlled amplifier which applies up to 80 dB of gain. The output of IFAMP 422 is then mixed in video detector (VDET) 426 with the video carrier, which has been extracted by video extraction filter (VCEF) 425 and video carrier limiter (VCLIM) 428 to produce the baseband video output.

The video carrier is compared by automatic frequency control (AFC) frequency detector (DET) 440 with a synthesized 20 MHz reference signal from AFCREF 442 to generate a frequency error signal which is passed back to LO1 for automatic frequency control.

The baseband video from VDET 426 is passed through a low-pass video bandwidth filter (VBBF) 430 to remove detection harmonics. The synchronization pulses are clamped to a reference level by sync clamp (SCMP) 432, which contains a phase-locked loop that is locked to the horizontal sync rate, RX LOCK 434, or vertical rate as an option, to control the clamp timing. Timing is also generated in SCMP 432 for AGC 438 and NINV 436. SCMP 432 amplifies the video to its output voltage level by a fixed gain of 20 dB. The clamped video is processed by noise inverter (NINV) 436 to remove excessively large positive or negative amplitude noise pulses.

The resulting video is tested for blanking to sync level by AGC 438, which adjusts the gain of LNTA 406 and IFAMP 422 ~~420~~ to maintain a standard 1 Volt peak-to-peak video signal. The video at this stage is buffered by video buffer (VBUF) 437 for off-chip drive using one of three signal impedance standards. VBUF 437 can also mute the signal.

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Audio processing is accomplished by extracting the audio signal at the output of IFAMP 422 through narrow bandpass audio IF filter (AIFF) 427. AIFF 427 produces an FM signal that is limited by audio IF limiter (AIFLIM) 444 to remove AM noise and distortion.

A delay-locked loop comprised of audio phase detector (APD) 455, audio delay-locked loop low-pass filter (ADLPF) 458, and audio phase shifter (APS) ~~audio delay line (ADLY)~~ 460 generates a signal that is delayed by 90° at the FM carrier frequency and mixes that signal with the limited FM signal in audio quadrature detector (AQD) 462 for quadrature detection of the FM. The audio baseband is passed through a 120 kHz low-pass audio baseband filter (ABF) 464 to remove detection

b

artifacts. The audio is buffered in audio buffer (ABUF) 466 for off-chip drive with selectable drive impedances at a 0.35 Volt (rms) level.

Control registers 471 and status registers 472 are interfaced to serial digital interface 170. Serial digital interface 170 in turn receives external commands from an external microprocessor or microcontroller through SERIAL DATA and SERIAL CLOCK control lines. In a preferred embodiment of the present invention, serial digital interface 170 may be a Inter Integrated Circuit (IIC or I^2C) interface, which is a proprietary specification of Philips Corporation.

System Block Specifications

The overall design of TV tuner 100 is guided by the limitations of what can be reliably designed on an integrated circuit. Blocks of TV tuner 100 are specified by referring to actual circuit elements so that capabilities such as noise, maximum signal handling, intermodulation and gain can be ascertained. Perhaps, the most important element of such a design is to limit expected chip performance so that it can be manufactured with a high yield and, therefore, a low cost. Note that all measurements in dBm are referenced to a standard video RF and baseband impedance of 75Ω.

RF Low-Pass Filter (RFLPF)

Table 1 lists the operating parameters of RF low-pass filter (RFLPF) 404 in accordance with a preferred embodiment of the present invention. RFLPF 404 is a 900 MHz third order low-pass filter comprising package lead inductance, printed circuit capacitance and on-chip capacitors. For most applications RFLPF

404 should be satisfactory for image (2400 - 3300 MHz) rejection in MIX1. External filtering matched to 75 ohm impedance could be added for extreme situations, such as very large out-of-band signals.

Table 1: RFLPF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	-1		0	dB
f_c	Cutoff Frequency (-3 dB)		900		MHz

Low Noise Transconductance Amplifier (LNTA)

Table 2 lists the operating parameters of low-noise transconductance amplifier (LNTA) 406 in accordance with a preferred embodiment of the present invention. LNTA 406 is the front end of the receiver. The single-ended input signal is converted to a differential current which will be passed through a mixer chain. The critical function of this circuit is to maintain a very low noise figure (*NF*), a substantial gain (G_V), and a high third-order intermodulation intercept (IP_3). The gain is determined relative to the rest of the circuit so that the noise figure of this block dominates the total receiver noise figure. Note that, since this is a transconductance amplifier, the gain is defined by the load impedance of the final mixer stage and must be properly defined before a final determination of LNTA 406 performance. AGC capability is included to reduce the impact of distortion, such as intermodulation or crossmodulation, in the presence of high level signals. The input is impedance matched to 75Ω.

Table 2: LNTA Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain - no AGC	20		30	dB
G_{VAGC}	Voltage Gain - full AGC	-10		0	dB
NF	Noise Figure	4.5		5.4	dB
P_{-1dB}	1 dB Gain Compression Input Level	-24			dBm
IP_3	Third Order Input Intercept	-10			dBm
f_{in}	Input Frequency	40		900	MHz
I_{DC}	DC Bias Current		20		mA

First Mixer (MIX1)

Table 3 lists the operating parameters of first mixer (MIX1) 408 in accordance with a preferred embodiment of the present invention. Proper rejection of any undesired images and spurious signals due to local oscillator harmonics requires that MIX1 be configured for an up-conversion scheme with high-side injection. Thus, the output of MIX1 for the visual carrier is at the microwave frequency of 1200 MHz, the audio carrier is at 1195.5 MHz, and local oscillator 450 input runs between 1.2 and 2.1 GHz. A single-pole low-pass element is designed into the output circuitry of MIX1 to assist in the rejection of any down-converted signals due to local oscillator harmonics. The noise and distortion effects of MIX1 are included in LNTA 406 specifications. The design will include a bond-wire parallel LC network option to further band-limit any image energy at the output of this mixer if necessary.

Table 3: MIX1 Specifications

Parameter	Description	Min	Typ	Max	Unit
f_{RF}	RF Input Frequency	40		900	MHz
f_{LO}	LO1 Input Frequency	1200		2100	MHz
f_{IF}	IF Output Frequency	1195		1205	MHz
V_{LO}	LO1 Input Voltage		0.2		V(rms)

Second Mixer (MIX2)

Table 4 lists the operating parameters of second mixer (MIX2) 410 in accordance with a preferred embodiment of the present invention. The first IF of 1.2 GHz (visual carrier) is input to MIX2, which is an image-rejection mixer, which down-converts to a 20 MHz second IF. The image rejection is done to negate the need for filtering of the first mixer output in order to reject signals falling on the image at 1160 MHz from being down-converted by MIX2 if MIX2 were just an ordinary subtractive mixer. An important feature of MIX2 is that it must achieve a high image rejection (at least 50dB) without the need for post-fabrication

InsB2 B2 ~~trimming techniques to calibrate MIX2. The 50dB image rejection may be achieved by paralleling two Gilbert cell multipliers at MIX2 input and using fairly high-order balanced RC-phase shifters with very low~~ component sensitivity with resulting phase errors via summer 615 of

Ins C1 C1 ~~less than 0.2 degrees over a 20-MHz bandwidth. These are traditionally avoided due to relatively high noise~~ levels, however, in a preferred embodiment of TV tuner 100, the previous stages provide sufficient gain to reduce the noise impact.

Table 4: MIX2 Specifications

Parameter	Description	Min	Typ	Max	Unit
G_I	Current Gain		0		dB
NF	Noise Figure			10	dB
f_{RF}	RF Input Frequency	1195		1205	MHz
f_{LO}	LO2 Input Frequency		1180		MHz
f_{IF}	IF Output Frequency	15.25		21.25	MHz
V_{LO}	LO2 Input Voltage		0.2		V(rms)
A_{IM}	Input Image Rejection	60			dB

Second IF Filter (SIFF)

Table 5 lists the operating parameters of second IF filter (SIFF) 420 in accordance with a preferred embodiment of the present invention. The 20 MHz IF signal is filtered for channel bandwidth in this integrated transconductor-capacitor (G_m-C) filter. A transitional Gaussian bandpass filter is used to provide adjacent channel and noise rejection. SIFF 420 keeps the distortions due to noise and out-of-band signals below -54 dBc. The in-channel group delay is to be maximally flat to the color subcarrier frequency. The aural signal passes through SIFF 420. Since the up-conversion scheme inverts the channel frequencies at the output of MIX2, the visual carrier is now at 20MHz and the aural carrier is at 15.5 MHz. The adjacent visual carrier is now at 14 MHz and the adjacent aural carrier is at 21.5 MHz leading to the filter specifications given below.

Table 5: SIFF Specifications

Parameter	Description	Min	Typ	Max	Unit
f_{pl}	Lower Passband Limit $A(\omega)<0.5$ dB		16.75		MHz
f_{ph}	Upper Passband Limit $A(\omega)<0.5$ dB		19.25		MHz
f_{sl}	Lower Stopband Limit $A(\omega)>54$ dB		15.25		MHz
f_{sh}	Upper Stopband Limit $A(\omega)>54$ dB		20.75		MHz
G_V	Voltage Gain		0		dB
IP_3	Third Order Input Intercept	20			dBm
NF	Noise FIgure			15	dB

IF Amplifier (IFAMP)

Table 6 lists the operating parameters of IF amplifier (IFAMP) 422 in accordance with a preferred embodiment of the present invention. The 20MHz signal is amplified by 80 dB in this block. The gain can be reduced to 0 dB by application of an automatic gain control (AGC) signal voltage. The AGC is implemented in fully differential form for best on-chip noise rejection.

Table 6: IFAMP Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain - No AGC		80		dB
G_V	Voltage Gain - Full AGC		20		dB
NF	Noise Figure			10	dB
BW	Bandwidth	25			MHz

Visual Carrier Extraction Filter (VCEF)

Table 7 lists the operating parameters of visual carrier extraction filter (VCEF) 425 in accordance with a preferred embodiment of the present invention. The video carrier at 15 MHz is extracted from IFAMP 422 output for synchronous detection processing. VCEF 425 passes along any incidental carrier phase modulation (ICPM) in the Nyquist filter without any group delay distortion for best detector performance.

Table 7: VCEF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_0	Center Frequency		20		MHz
BW	Bandwidth	0.5	1	3	MHz
NF	Noise Figure			5	dB

Visual Carrier Limiter (VCLIM)

Table 8 lists the operating parameters of visual carrier limiter (VCLIM) 428 in accordance with a preferred embodiment of the present invention. The synchronous detector requires a fixed amplitude reference carrier. This is derived via VCEF 425 and then passed through VCLIM 428. With input level variations of 20 dB and depth of modulation possibly exceeding 95 percent, VCLIM 428 output is stable within an accuracy of -50 dB. The circuit delay of VCEF 425 and VCLIM 428 combined must not exceed 45 degrees at the visual carrier frequency so that the detector output level and noise behavior is not degraded.

Table 8: VCLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	46			dB
V_{th}	Limiter Input Threshold		10		mV(p-p)
V_{limit}	Limiter Output Level	2			V(p-p)
NF	Noise Figure			5	dB

Video Detector (VDET)

Table 9 lists the operating parameters of video detector (VDET) 426 in accordance with a preferred embodiment of the present invention. Video is detected synchronously by applying the output of IFAMP 422 as a first input signal and the output of VCLIM 428 as a second input signal to a double-balanced mixer (VDET 426). The output of VDET 426 is configured with a single low-pass pole set to allow 4.2 MHz to pass with a 0.5 dB loss.

Table 9: VDET Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
BW	Output -0.5 dB Bandwidth		4.2		MHz
V_{video}	Video Output Level		0.1		V(p-p)
NF	Noise Figure			5	dB

Video Baseband Filter (VBBF)

Table 10 lists the operating parameters of video baseband filter (VBBF) 430 in accordance with a preferred embodiment of the present invention. The detected video is band-limited to 4.2 MHz by a multiple pole transitional Gaussian response low-pass filter

which has been predistorted for both compensation of SIFF 420 response and group delay characteristics conforming to FCC regulations. VBBF 430 rejects the video carrier at 20 MHz and the high level 40 MHz modulated signal at the output of VDET 426 with minimum pass-band transient distortion.

Table 10: VBBF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_p	Passband Corner Frequency A(ω)<0.5 dB		4.2		MHz
f_s	Stopband Corner Frequency A(ω)>54 dB		8.0		MHz
NF	Noise Figure			5	dB

<u>Sync Clamp (SCMP)</u>

Table 11 lists the operating parameters of sync clamp (SCMP) 432 in accordance with a preferred embodiment of the present invention. The sync pulses are clamped via two techniques. A diode clamp is used to define the minimum level to which the sync tip is set. Once this is achieved, after one sync pulse, a sync slicer regenerates the sync pulses only and passes these on to a phase-locked loop (PLL). The PLL acquires the horizontal sync rate and generates a sync gating pulse approximately centered in the horizontal sync pulse.

Once the PLL has achieved lock, a gated clamp circuit using the sync gating pulse is enabled to define the sync level and the diode clamp is disabled. For processing of scrambled video, a line counter is inserted into the phase-locked loop to count a field, the sync is integrated for detection of vertical sync

and only the vertical sync pulse is used for the clamp gate. The sync gating pulse is used via a clocked delay of 5 microseconds to generate a blanking gate used for AGC processing. SCMP 432 also amplifies the video to its final output level. The output signal RX LOCK 434 becomes active when the phase-locked loop has achieved lock to the sync pulses.

Table 11: SCMP Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		20		dB
V_{Video}	Video Output Level		1		V(p-p)
f_{PLLH}	PLL Operating Frequency H-Sync Mode		15750		kHz
f_{PLLV}	PLL Operating Frequency V-Sync Mode		60		Hz

Noise Inverter (NINV)

Table 12 lists the operating parameters of noise inverter (NINV) 436 in accordance with a preferred embodiment of the present invention. SCMP 432 video output will include extreme transient noise pulses in both black and white directions. The black noise is typically a large RF transient due to some form of interference. The signal will drop below the blanking level during the horizontal line picture period and may upset downstream sync detect circuits. A comparator detects these levels below the blanking level during active picture, or non-blanked, periods and enables an inverting video path during the event which is summed with the original video. The summation cancels the effect of the noise pulse so that the sync level region is not disturbed. Extreme positive excursions, which appear as white, are also cancelled when they exceed

the equivalent 108 IRE white level, or less than 5 percent modulation. These typically are due to desensitization effects of the receiver when pulse interference exists on an out of band frequency and typically result in the disruption of the synchronous detector. Positive noise inversion is accomplished using a comparator referenced to the 108 IRE level and enabling a video inversion path to a summer as for the negative case above.

Table 12: NINV Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p-p)
V_{WINS}	White Noise Inverter Insertion Level		95		IRE
V_{BINS}	Black Noise Inverter Insertion Level		10		IRE
V_{WTH}	White Inverter Threshold		108		IRE
V_{BTH}	Black Inverter Threshold Non-Blanked		0		IRE

Automatic Gain Control (AGC)

Table 13 lists the operating parameters of automatic gain control (AGC) 438 in accordance with a preferred embodiment of the present invention. The clamped video is fed into a sample-and-hold circuit controlled by the blanking gate. The sample-and-hold circuit output drives a differential amplifier which has a reference level at its other input equal to the desired output blanking level for the standard NTSC video waveform.

On each horizontal sync pulse the output of the differential amplifier represents the gain error of the receiver to achieve the video waveform. This is fed

back to LNTA 406 and IFAMP 422 to adjust the gain as necessary. The timing and gain of the differential amplifier must be set for fast transient response time to account for aircraft induced flutter (fading) and unconditional stability. The response can be changed to update gain only during the vertical sync period, just as with SCMP 432, so that scrambled video may pass through the tuner without any adverse effects.

The AGC voltage, defined by the performance of the RF and IF amplification behavior and input signal level, is measured via 4 bit analog-to-digital converter and stored as the data word Received Signal Strength Indicator (RSSI) 439. Automatic gain control is applied first to IFAMP 422 to reduce the video level, and beyond 50 dB of reduction, also to LNTA 406, which implements a delayed AGC function.

Table 13: AGC Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p-p)
V_{s-b}	Input Sync to Blanking Level Reference		0.286		V
G_{AGC}	AGC Loop Gain	70			dB
GBW_H	AGC Gain-Bandwidth H-Mode		150		Hz
GBW_V	AGC Gain-Bandwidth V-Mode		6		Hz
Φ_{MAGC}	AGC Loop Phase Margin	60			°

Video Output Buffer (VBUF)

Table 14 lists the operating parameters of video buffer (VBUF) 437 in accordance with a preferred embodiment of the present invention. The video is at an internal level of 1 volt peak-to-peak differential.

VBUF 437 converts this signal to a current via a linear transconductance amplifier that drives a load resistor. Feedback is arranged via an external signal to allow for continuous adjustment of the gain and offset of the amplifier so that no AC coupling is required in the next stage. An on-chip integrator is used to define the feedback level and reject the video signal. The control signal VOUT MODE 445 enables on-chip loads of 75Ω, 1000Ω or no load, which equates to a current source output. The control signal VIDEO MUTE 446 enables a video mute switch which sets the output to the blanking level.

Table 14: VBUF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Video Input Level		1		V(p-p)
V_{OV}	Video Output Level (R_L>70 Ω)		1		V(p-p)
V_{OVR}	Video Output DC Reference (R_L>70Ω)	0.5		3.5	V
Z_{OUT}	Output Impedance 75Ω Mode	73	75	77	Ω
Z_{OUT}	Output Impedance 1000Ω Mode	980	1000	1020	Ω
Z_{OUT}	Output Impedance Current Mode		10		kΩ

Automatic Frequency Control (AFC) Frequency Detector

Table 15 lists the operating parameters of automatic frequency control (AFC) frequency detector (DET) 440 in accordance with a preferred embodiment of the present invention. The extracted IF video carrier is compared to a fixed reference at 20 MHz and used to drive LO1 in a direction to equalize the two. The error in LO1 from its frequency control setting is the

signal AFC ERROR 452. The 20 MHz reference is generated via a frequency synthesizer from the master reference. The 20 MHz reference is compared to the extracted video carrier signal at the output of VCEF 425 using a digital frequency comparator which drives a seven bit up-down counter that is summed with the low bits of the frequency code sent to LO1 from the I^2C interface 170 ~~470~~. Thus, the frequency error appears in the counter which can be read via the interface bus. This counter is set to 1000000b when there is no frequency error. This condition sets the status signal FREQ LOCK 453.

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Table 15: AFC Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{CIN}	Video Carrier Input Level		0.1		V(p-p)
f_{ERROR}	Frequency Error From Nominal	-500		500	kHz
r_{UPDATE}	Frequency Update Rate			30	Hz

AFC Reference Oscillator (AFCREF)

Table 16 lists the operating parameters of AFC reference oscillator (AFCREF) 442 in accordance with a preferred embodiment of the present invention. AFCREF 442 is a fixed frequency synthesizer running at 20 MHz, which is phase-locked to the quartz crystal frequency reference. Phase noise is not particularly critical due to the application of this circuit at low modulation frequencies only.

Table 16: AFCREF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p-p)

Parameter	Description	Min	Typ	Max	Unit
N_{op}	Phase Noise for 20 Hz< f_m< 1 MHz			-60	dBc
f_0	Operating Frequency		20		MHz

First Local Oscillator (LO1)

Table 17 lists the operating parameters of first local oscillator (LO1) 450 in accordance with a preferred embodiment of the present invention. LO1 is a phase-locked frequency synthesizer configured for specific levels of phase noise to allow for best receiver sensitivity without excessive spurious noise responses. It is digitally controlled and utilizes a quartz crystal reference for frequency and in-band phase noise control. The frequency control input FREQUENCY 451 from the serial digital interface 170 bus has sixteen (16) bits with 62.5 kHz LSB resolution. Four (4) more bits below the LSB as well as the first 3 LSB's are used by the AFC for fine tuning.

Table 17: LO1 Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p-p)
N_{op}	Phase Noise for f_m>1 MHz			-110	dBc
N_{op}	Phase Noise for 60 Hz < f_m< 1 MHz			-80	dBc
f_0	Operating Frequency	1200		2100	MHz
f_{STEP}	Frequency Step		3.906		kHz
f_{REF}	Reference Frequency		62.5		kHz

Second Local Oscillator (LO2)

Table 18 lists the operating parameters of second local oscillator (LO2) 412 in accordance with a preferred embodiment of the present invention. LO2 is a fixed tuned frequency synthesizer set to 1180 MHz. It uses the same frequency reference as LO1 and has similar phase noise characteristics.

Table 18: LO2 Specifications

Parameter	Description	Min	Typ	Max	Unit
V_C	Carrier Level		0.2		V(p-p)
N_{op}	Phase Noise for f_m>1 MHz			-110	dBc
N_{op}	Phase Noise for 60 Hz < f_m< 1 MHz			-80	dBc
f_0	Operating Frequency		1180		MHz
f_{REF}	Reference Frequency		62.5		kHz

Audio IF Filter (AIFF)

Table 19 lists the operating parameters of audio IF filter (AIFF) 427 in accordance with a preferred embodiment of the present invention. The audio IF signal is tapped from the output of IFAMP 422 and run through a narrow-band filter operating a 15.5 MHz with a bandwidth of 300 kHz. AIFF 427 rejects the video carrier and the burst frequency components.

Table 19: AIFF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
f_0	Center Frequency		15.5		MHz
BW	Bandwidth		300		kHz
NF	Noise Figure			10	dB

TUNE 048200

Audio IF Limiter (AIFLIM)

Table 20 lists the operating parameters of audio IF limiter (AIFLIM) 444 in accordance with a preferred embodiment of the present invention. The audio IF is limited with a 40 dB possible variation in input level based on AGC 438 and the difference between sources of the broadcast signal. The output level accuracy over this range is better than -40 dB.

Table 20: AIFLIM Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain	40			dB
V_{th}	Limiter Input Threshold		2		mV(p-p)
V_{limit}	Limiter Output Level		0.2		V(p-p)
NF	Noise Figure			10	dB

Audio Phase Detector (APD)

Table 21 lists the operating parameters of audio phase detector (APD) 455 in accordance with a preferred embodiment of the present invention. The audio IF is compared with APS 460 output to generate a phase error signal to drive APS 460 in a delay-locked loop. APD 455 is a phase detector configured for fast loop acquisition and minimum phase error once acquired.

Table 21: APD Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{IN}	Voltage Input Level		0.35		V(rms)

Audio DLL Low Pass Filter (ADLPF)

Table 22 lists the operating parameters of audio delay-locked loop low-pass filter (ADLPF) 458 in accordance with a preferred embodiment of the present invention. ADLPF 458 must not respond to any frequencies above 20 Hz in order to achieve the full audio bandwidth in the detected signal. The low-pass filter accommodates this requirement.

Table 22: APLPF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_V	Voltage Gain		0		dB
BW	Loop Bandwidth			20	Hz
Z_L	Closed Loop Zero		10		Hz

Audio Phase Shifter (APS)

Table 23 lists the operating parameters of audio phase shifter (APS) 460 in accordance with a preferred embodiment of the present invention. APS 460 is a voltage controlled phase shifter which is locked to the audio carrier via a delay-locked loop. APS 460 generates a 90° degree phase shift in the output signal at the aural carrier frequency of 15.5 MHz.

Table 23: AQO Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{OUT}	Output Voltage		0.1		V
f_0	Center Frequency		15.5		MHz
0_S	Phase Shift		90		Deg.

Audio Quadrature Detector (AQD)

Table 24 lists the operating parameters of audio quadrature detector (AQD) 462 in accordance with a preferred embodiment of the present invention. AQD 462 is a mixer that is used to compare the incoming audio IF signal with a delayed version of the IF signal running in quadrature with the incoming carrier. The output is the desired audio baseband signal.

Table 24: AQD Specifications

Parameter	Description	Min	Typ	Max	Unit
G_v	Voltage Gain		0		dB

Audio Baseband Filter (ABF)

Table 25 lists the operating parameters of audio baseband filter (ABF) 464 in accordance with a preferred embodiment of the present invention. ABF 464 is a high-order Chebychev low-pass design with a cutoff at 120 kHz. This is wide enough to pass all multiplex audio subcarriers for external decoding.

Table 25: ABF Specifications

Parameter	Description	Min	Typ	Max	Unit
G_v	Voltage Gain		10		dB
f_p	Passband Corner Frequency A(ω)<0.5 dB		120		kHz
f_s	Stopband Corner Frequency A(ω)>54 dB		300		kHz

Audio Output Buffer (ABUF)

Table 26 lists the operating parameters of audio output buffer (ABUF) 466 in accordance with a preferred

embodiment of the present invention. The audio is at an internal level of 0.1 volts peak to peak. ABUF 466 converts this signal to a current via a linear transconductance amplifier which then drives a load resistor. Feedback is arranged via an external signal to allow for continuous adjustment of the gain and offset of the amplifier so that no AC coupling is required to the next stage. An on-chip integrator is used to define the feedback level and reject the audio signal.

The control signal AOUT MODE 467 enables on-chip loads of 600Ω, 10 kΩ or no load, which equates to a current source output. The control signal AUDIO MUTE 468 controls an audio mute switch which disables the output.

Table 26: ABUF Specifications

Parameter	Description	Min	Typ	Max	Unit
V_{Video}	Audio Input Level		0.1		V(p-p)
V_{OV}	Audio Output Level (R_L>500 Ω)		0.35		V(p-p)
V_{OVR}	Audio Output DC Reference (R_L>500 Ω)	0.5		3.5	V
Z_{OUT}	Output Impedance 600 Ω Mode	588	600	612	Ω
Z_{OUT}	Output Impedance 10 kΩ Mode	9.8	10.0	10.2	kΩ
Z_{OUT}	Output Impedance Current Mode		100		kΩ

Serial Digital Interface

Tables 27-29 define the registers, register bits and bit functions that are addressable by serial digital interface 170 in accordance with a preferred embodiment of the present invention. For example,

control and status may be performed via an Inter Integrated Circuit (I^2C) bus interface. Serial digital interface 170 contains all registers and provides access to all parallel digital on-chip functions via the serial bus.

Table 27: Register Definition

Register Name	Description	Register Address	Read/Write
FREQ0	Frequency - Low byte	00	Read/Write
FREQ1	Frequency - High byte	01	Read/Write
CTRL0	Functional control	02	Read/Write
CTRL1	Functional control	03	Read/Write
STAT0	Functional status	04	Read
STAT1	Functional status	05	Read
STAT2	Functional status	06	Read
CMPY0	Company code - Low byte	07	Read
CMPY1	Company code - High byte	08	Read
REV0	Revision code - Low byte	09	Read
REV1	Revision code - High byte	0A	Read

The definitions of the bits used in each register are:

Table 28: Register Bit Definitions

Register	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2	Bit 1	Bit 0
FREQ0	F7	F6	F5	F4	F3	F2	F1	F0
FREQ1	F15	F14	F13	F12	F11	F10	F9	F8
CTRL0	PWRDN	AMUTE	VMUTE		MODE3	MODE2	MODE1	MODE0
CTRL1					VRS1	VRS0	ARS1	ARS0
STAT0	RLOCK	FLOCK				AFC2	AFC1	AFC0

Register	Bit 7	Bit 6	Bit 5	Bit 4	Bit 3	Bit 2	Bit 1	Bit 0
STAT1					RSSI3	RSSI2	RSSI1	RSSI0
STAT2	T7	T6	T5	T4	T3	T2	T1	T0
CMPY0	C7	C6	C5	C4	C3	C2	C1	C0
CMPY1	C15	C14	C13	C12	C11	C10	C9	C8
REV0	R7	R6	R5	R4	R3	R2	R1	R0
REV1	R15	R14	R13	R12	R11	R10	R9	R8

The status registers are read only. Attempting to write a status register will have no effect. The function of each bit is defined as:

Table 29: Bit Functions

Bit Name	Function	Description
F0-15	Receiver frequency	$f=0.0625 \cdot F+k_f$ (MHz) (F15=MSB, F0=LSB)
PWRDN	Circuit power-down	When high the receiver is disabled, the video and audio outputs are muted and the I^2C remains functional. The circuit draws minimum current.
AMUTE	Audio mute	Audio output is disabled when high.
VMUTE	Video mute	Video output is disabled when high.
MODE0-3	Receiver mode select	Selects receiver operating mode (unused, reserved for NTSC/PAL/SECAM switching).
VRS0-1	Video output impedance select	00=75Ω, 01=1 kΩ, 02=Current Source.
ARS0-1	Audio output impedance select	00=600Ω, 01=10 kΩ, 02=Current Source.
RLOCK	Receiver lock	High if AFC is locked to incoming signal
FLOCK	Synthesizer lock	High if frequency synthesizer locked
AFC0-2	AFC offset	Indicates magnitude and direction of mistuning of receiver. Three MSB's of AFC counter.

Bit Name	Function	Description
RSSI0-3	Received signal strength indicator	Indicates relative signal strength of signal. 0000 is no effective signal, 1111 is large signal. (RSSI3=MSB, RSSI0=LSB)
T0-7	Test data	Data used for test purposes only.
C0-15	Company code	Unique identifier for Cirrus Logic I^2C parts
R0-15	Revision code	Unique identifier for part number and revision

Bias and Control (BC)

Table 30 lists the operating parameters of bias and control logic (BC) 180 in accordance with a preferred embodiment of the present invention. Bias voltages and currents are generated to define specific levels need for each circuit element in the present invention. The control signal POWERDOWN 476 controls the power-down function via BC 180.

Table 30: BC Specifications

Parameter	Description	Min	Typ	Max	Unit
F_{osc}	Crystal Oscillator Frequency Reference		8		MHz

Although the present invention and its advantages have been described in detail, it should be understood that various changes, substitutions and alterations can be made herein without departing from the spirit and scope of the invention as defined by the appended claims.

(2836-P223US) PATENT

WHAT IS CLAIMED IS:

~~1. A television receiver comprising:~~

~~a receiver input coupled to an RF signal source;~~

~~a first reference signal having a first operating frequency;~~

~~a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;~~

~~a second reference signal having a second operating frequency; and~~

~~a second mixer having a first input coupled directly to an output of said first mixer and a second input coupled to said second reference signal.~~

2. The television receiver as set forth in Claim 1 wherein said first mixer is a subtractive mixer.

~~3. The television receiver as set forth in Claim 2 wherein said ~~second~~ mixer is an image rejection mixer.~~

3 ~~4~~. The television receiver as set forth in Claim 2 wherein a minimum frequency value of said first reference signal is greater than an input cutoff frequency of an RF signal received from said RF signal source.

4 ~~5~~. The television receiver as set forth in Claim 1 wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

~~6.~~ 5. The television receiver as set forth in Claim ~~5~~ 4 wherein said first mixer subtractively mixes said first reference signal and an RF signal received from said RF signal source to thereby generate a first IF signal on said first mixer output.

~~7.~~ 6. The television receiver as set forth in Claim ~~6~~ 5 wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

~~8.~~ 7. The television receiver as set forth in Claim ~~7~~ 6 wherein said first predetermined frequency is 1200 Megahertz.

~~9.~~ 8. The television receiver as set forth in Claim ~~7~~ 6 wherein said selected carrier frequency is selected by varying said first operating frequency.

D

~~10.~~ 9. The television receiver as set forth in Claim ~~6~~ 1 wherein said first mixer is coupled to said receiver input by an input filter which filters out all frequency components in said RF signal above an input cutoff frequency.

~~11.~~ 10. The television receiver as set forth in Claim ~~10~~ 9 wherein said minimum frequency value of said first reference signal is greater than said input cutoff frequency.

Sub a4

12. The television receiver as set forth in Claim 7 wherein said second mixer mixes said second reference signal and said first IF signal to thereby generate a second IF signal.

12 ~~13~~. The television receiver as set forth in Claim 11 ~~12~~ wherein said second operating frequency is fixed.

13 ~~14~~. The television receiver as set forth in Claim 12 ~~13~~ wherein said second mixer rejects from said second IF signal an image signal of said second reference signal.

14 ~~15~~. The television receiver as set forth in Claim 13 ~~14~~ wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

15 ~~16~~. The television receiver as set forth in Claim 13 ~~14~~ wherein said minimum frequency value of said first reference signal is greater than an input cutoff frequency of said RF signal.

Sub a5 ~~17. The television receiver as set forth in Claim 1 further comprising a first IF signal generated by said first mixer wherein said second mixer mixes said second reference signal and said first IF signal to generate a second IF signal.~~

17 ~~18~~. The television receiver as set forth in Claim 16 ~~17~~ wherein said second operating frequency is fixed.

18 ~~19~~. The television receiver as set forth in Claim 17 ~~18~~ wherein said second mixer rejects from said second IF signal an image signal of said first IF signal.

19 ~~20~~. The television receiver as set forth in Claim 18 ~~19~~ wherein a selected carrier frequency in an RF signal received from said RF signal source appears at a first

predetermined frequency in said first IF signal and at a second predetermined frequency in said second IF signal.

20 ~~21~~. The television receiver as set forth in Claim 19 ~~20~~ wherein said second predetermined frequency is a ~~the~~ difference in frequency between said first predetermined frequency and said second operating frequency.

21 ~~22~~. The television receiver as set forth in Claim 20 ~~21~~ wherein said second operating frequency is less than said first predetermined frequency and the frequency of said image signal rejected by said second mixer is less than said second operating frequency by an amount equal to said second predetermined frequency.

22 ~~23~~. The television receiver as set forth in Claim 20 ~~21~~ wherein said first predetermined frequency is 1200 Megahertz, said second operating frequency is 1180 Megahertz, and said second predetermined frequency is 20 Megahertz.

Sub 26

24. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value; and

a second mixer having a first input coupled to an output of said first mixer and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

23 24 ~~25~~. The television receiver as set forth in Claim ~~24~~ wherein said minimum frequency value is greater than an input cutoff frequency of an RF signal received from said RF signal source.

24 25 ~~26~~. The television receiver as set forth in Claim ~~25~~ wherein said first mixer subtractively mixes said RF signal and said first reference signal to thereby produce a first IF signal, wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal and said first predetermined frequency is greater than said selected carrier frequency.

25 26 ~~27~~. The television receiver as set forth in Claim ~~26~~ wherein said second mixer mixes said first IF signal and said second reference signal to thereby produce a second IF signal, wherein said selected carrier frequency appears in said second IF signal at a second

predetermined frequency and said second predetermined frequency is less than said first predetermined frequency and said second operating frequency.

27 ~~28~~. The television receiver as set forth in Claim 26 ~~27~~ wherein said second mixer rejects at least one image signal associated with said second operating frequency from said second IF signal.

Sub a7

29. A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

~~30~~ 29. The method as set forth in Claim ~~29~~ 28 including the further step of filtering out of the RF signal all frequency components above an input cutoff frequency.

~~31~~ 30. The method as set forth in Claim ~~30~~ 29 wherein the first operating frequency is greater than the input cutoff frequency.

~~32~~ 31. The method as set forth in Claim ~~31~~ 30 wherein the step of mixing the RF signal with the first reference signal is a subtractive mixing of the RF signal and the first reference signal.

~~33~~ 32. The method as set forth in Claim ~~32~~ 31 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a frequency higher than the RF signal.

~~34~~ 33. The method as set forth in Claim ~~29~~ 28 wherein the step of mixing the first IF signal with the second reference signal further includes the substep of removing from the second IF signal at least one image signal associated with the ~~second~~ first IF signal.

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Sub a8

35. A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

35 ~~36~~. The method as set forth in Claim 34 ~~35~~ wherein the first operating frequency is greater than the input cutoff frequency.

36 ~~37~~. The method as set forth in Claim 35 ~~36~~ wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in the RF signal appears in the first IF signal at a first predetermined frequency and wherein the first predetermined frequency is higher than the RF signal.

37 ~~38~~. The method as set forth in Claim 34 ~~35~~ wherein the step of mixing the first IF signal with the second reference signal further includes the substep of removing from the second IF signal at least one image signal associated with the first ~~second~~ IF signal.

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SUB B6

39. ~~The method~~ as set forth in Claim 35 wherein the step of mixing the ~~first IF signal~~ with the second

~~reference signal is a down conversion mixing, wherein the selected carrier signal appearing at the first predetermined frequency appears in the second IF signal at a second predetermined frequency and wherein the second predetermined frequency is lower than the first IF signal.~~

39 ~~40~~. The method as set forth in Claim 34 ~~35~~ including the further step of filtering the second IF signal with a bandpass filter.

40 ~~41~~. The method as set forth in Claim 39 ~~40~~ wherein the bandpass filter has a passband between 15 Megahertz and 21 Megahertz.

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41 ~~42~~. The method as set forth in Claim 39 ~~40~~ ~~35~~ wherein the bandpass filter is a Nyquist slope filter having a cutoff frequency of 20 Megahertz.

Sub a9 43. A television receiver comprising:

an input filter coupled to an RF signal source;

a first amplifier coupled to an output of said input filter;

a first mixer having a first input and a second input, wherein said first input of said first mixer is coupled to an output of said first amplifier;

a first oscillator coupled to said second input of said first mixer, wherein said first oscillator generates a first reference signal having a first operating frequency;

a second mixer having a first input and a second input, wherein said first input of said second mixer is coupled to an output of said first mixer;

a second oscillator coupled to said second input of said second mixer, wherein said second oscillator generates a second reference signal having a second operating frequency; and

a second filter coupled to an output of said second mixer.

43 ~~44.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said input filter is a low-pass filter having a -3dB cutoff frequency of 900 Megahertz.

44 ~~45.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said first amplifier is a transconductance amplifier having a noise figure between 4.5 dB and 5.4 dB.

45 ~~46.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein the gain of said first amplifier is

automatically controlled by an automatic gain control circuit of said television receiver.

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46 ~~47.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said first mixer is an up-converter which subtractively mixes a filtered and amplified RF signal having a maximum cutoff frequency with said first reference signal to thereby generate a first IF signal, wherein a lower limit of said first operating frequency is greater than said maximum cutoff frequency.

47 ~~48.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said first oscillator is a phase-locked synthesizer and said first operating frequency has an upper limit of 2100 Megahertz and a lower limit of 1200 Megahertz.

B

48 ~~49.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said second mixer is an image-rejection mixer which down-converts a first IF signal outputted ~~received~~ from said first mixer to thereby generate a second IF signal.

49 ~~50.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said second oscillator is a fixed tuned-frequency synthesizer and said second operating frequency is approximately 1180 Megahertz.

50 ~~51.~~ The television receiver as set forth in Claim 42 ~~43~~ wherein said second filter is a bandpass filter having an upper passband limit of 20.75 Megahertz and a lower passband limit of 15.8 Megahertz.

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42 51 ~~43~~ ~~52~~. The television receiver as set forth in Claim ~~51~~ wherein said second filter is a Nyquist slope filter.

B

51 52 ~~53~~. The television receiver as set forth in Claim ~~52~~ wherein said second filter has a -6 dB gain point at ~~the~~ a visual carrier frequency of an IF signal generated by said second mixer.

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52 53 ~~54~~. The television receiver as set forth in Claim ~~53~~ wherein said ~~video~~ visual carrier frequency is approximately 20 Megahertz and said second filter has a passband corner frequency of approximately 19.25 Megahertz.

52 54 ~~55~~. The television receiver as set forth in Claim ~~53~~ wherein said second filter is automatically tuned by measuring the level of synchronization pulses on a input of said second filter and on an output of said second filter and adjusting a cut-off frequency of said second filter so that there is a 6 dB loss on said output of said second filter.

55 ~~56~~. The television receiver as set forth in Claim 42 ~~43~~ wherein said television receiver further comprises:

a second amplifier coupled to an output of said second filter.

56 ~~57~~. The television receiver as set forth in Claim 55 ~~56~~ wherein the gain of said second amplifier is automatically controlled by an automatic gain control circuit of said television receiver.

SUB B7

58. The television receiver as set forth in Claim 56 wherein said television receiver further comprises:

a fourth filter coupled to an output of said second amplifier, wherein said fourth filter is a bandpass filter for extracting a video carrier signal from an output signal of said second amplifier; and

a third mixer having a first input and a second input, wherein said first input of said third mixer is coupled to said output of said second amplifier and said second input of said third mixer is coupled to an output of said fourth filter and wherein an output signal on an output of said third mixer is a video baseband signal.

B

58 ~~59~~. The television receiver as set forth in Claim 57 ~~58~~ wherein said second input of said third mixer is coupled to said ~~fourth~~ third filter by amplitude limiting circuitry, wherein said amplitude limiting circuitry provides a fixed-amplitude reference carrier signal to said third mixer.

B

59 ~~60~~. The television receiver as set forth in Claim 57 ~~58~~ wherein said television receiver further comprises frequency detection circuitry having a first input and a second input, wherein said first input of said frequency detection circuitry is coupled to said output of said ~~fourth~~ third filter and said second input of said frequency detection circuitry is coupled to a third reference signal having a third operating frequency.

60 ~~61~~. The television receiver as set forth in Claim 59 ~~60~~ wherein said frequency detection circuitry compares the frequency of said video carrier signal to said third operating frequency to thereby produce an output

control signal, wherein said output control signal is used to control said first operating frequency of said first reference signal generated by said first oscillator.

61 ~~62~~. The television receiver as set forth in Claim 57 ~~58~~ wherein said television receiver further comprises a fourth ~~fifth~~ filter coupled to said output of said third mixer and wherein said fourth ~~fifth~~ filter is a low-pass filter that cuts off all signal components higher in frequency than said video baseband signal.

62 ~~63~~. The television receiver as set forth in Claim 61 ~~62~~ wherein said television receiver further includes a clamping circuit for clamping synchronization pulses in said video baseband signal from said fourth filter.

63 ~~64~~. The television receiver as set forth in Claim 62 ~~63~~ wherein said television receiver further comprises a noise inverter circuit coupled to said clamping circuit, wherein said noise inverter circuit removes transient noise pulses from said video baseband signal.

64 ~~65~~. The television receiver as set forth in Claim 63 ~~64~~ wherein said transient noise pulses comprise transient noise pulses that cause said video baseband signal to drop below a blanking level of said video baseband signal.

65 ~~66~~. The television receiver as set forth in Claim 63 ~~64~~ wherein said transient noise pulses comprise transient noise pulses that cause said video baseband signal to rise above a 108 IRE white level of said video baseband signal.

66 ~~67~~. The television receiver as set forth in Claim 57 ~~88~~ wherein said television receiver further comprises an automatic gain control circuit coupled to said output of said third mixer, wherein said automatic gain control circuit controls the gain of said second amplifier.

67 ~~68~~. The television receiver as set forth in Claim 66 ~~67~~ wherein said automatic gain control circuit controls the gain of said first amplifier.

68 ~~69~~. The television receiver as set forth in Claim 66 ~~67~~ wherein said automatic gain control circuit comprises a sample-and-hold circuit coupled to a first input of a differential amplifier.

69 ~~70~~. The television receiver as set forth in Claim 68 ~~69~~ wherein a second input of said differential amplifier is coupled to a reference voltage equal to an output blanking level of a standard NTSC video baseband signal.

70 ~~71~~. The television receiver as set forth in Claim 57 ~~58~~ wherein said television receiver further comprises a video output buffer amplifier coupled to said output of said third mixer.

71 ~~72~~. The television receiver as set forth in Claim 70 ~~71~~ wherein said video output buffer amplifier comprises a linear transconductance amplifier.

72 ~~73~~. The television receiver as set forth in Claim 71 ~~72~~ wherein said video output buffer amplifier further comprises blanking means for setting an output signal of said video output buffer amplifier to a video blanking level.

73 ~~74~~. The television receiver as set forth in Claim 72 ~~73~~ wherein said video output buffer amplifier further comprises switching means for varying an output load of said video output buffer amplifier.

SUB B8

~~75. The television receiver as set forth in Claim 52 wherein said television receiver further comprises a sixth filter coupled to an output of said second amplifier, wherein said sixth filter is a narrow-band filter for filtering an audio component signal from a video output signal on said output of said second amplifier.~~

BC

75
74 ~~76~~. The television receiver as set forth in Claim ~~75~~ wherein said ~~sixth~~ ~~fifth~~ third filter has a center frequency of 15.5 Megahertz and a band width of 300 Kilohertz.

Sub C6

B

B

~~77. The television receiver as set forth in Claim 75 wherein said television receiver further comprises:~~

~~audio phase detector circuitry coupled to an output of said ~~sixth~~ fifth filter; and~~

~~a fourth mixer having a first input and a second input, wherein said first input of said fourth mixer is coupled to said output of said ~~sixth~~ fifth filter and said second input of said fourth mixer is coupled to an output of said audio phase detector circuitry.~~

77
~~78~~. The television receiver as set forth in Claim 76 ~~77~~ wherein said audio phase detector circuitry comprises a delay-locked loop, wherein an output signal of said delay-locked loop operates in phase quadrature with a center frequency of said audio component signal on said output of said ~~sixth~~ ~~fifth~~ third filter.

BC

Ket

B

B

~~79. The television receiver as set forth in Claim 77 wherein said television receiver further comprises a ~~seventh~~ sixth filter coupled to an output of said fourth mixer, wherein said ~~seventh~~ sixth filter is a low-pass filter with a cut-off frequency of 120 Kilohertz.~~

79 ~~80.~~ The television receiver as set forth in Claim 76 ~~77~~ wherein said television receiver further comprises an audio output buffer amplifier coupled to an output of said third ~~fourth~~ mixer.

c

80 ~~81~~. The television receiver as set forth in Claim 79 ~~80~~ wherein said audio output buffer amplifier comprises a linear transconductance amplifier.

81 ~~82~~. The television receiver as set forth in Claim 79 ~~80~~ wherein said audio output buffer amplifier further comprises muting means for setting an output signal of said audio output buffer amplifier to zero volts.

82 ~~83~~. The television receiver as set forth in Claim 80 ~~80~~ wherein said audio output buffer amplifier further comprises switching means for varying an output load of said audio output buffer amplifier.

(2836-P223US) PATENT

Sub a10

84. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a first bandpass filter coupled to an output of said first mixer;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to said second reference signal.

84 ~~85~~. The television receiver as set forth in Claim 83 ~~84~~ wherein said first mixer is a subtractive mixer.

85 ~~86~~. The television receiver as set forth in Claim 84 ~~85~~ wherein a minimum frequency value of said first reference signal is greater than an input cutoff frequency of an RF signal outputted ~~received~~ from said RF signal source.

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86 ~~87~~. The television receiver as set forth in Claim 83 ~~84~~ wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

87 ~~88~~. The television receiver as set forth in Claim 86 ~~87~~ wherein said first mixer subtractively mixes said first reference signal and an RF signal outputted ~~received~~ from

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said RF signal source to thereby generate a first IF signal on said first mixer output.

88 ~~89~~. The television receiver as set forth in Claim 87 ~~88~~ wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

89 ~~90~~. The television receiver as set forth in Claim 88 ~~89~~ wherein said first predetermined frequency is 1200 Megahertz.

90 ~~91~~. The television receiver as set forth in Claim 88 ~~89~~ wherein said selected carrier frequency is selected by varying said first operating frequency.

91 ~~92~~. The television receiver as set forth in Claim 87 ~~88~~ wherein said first mixer is coupled to said receiver input by an input filter which filters out all frequency components in said RF signal above an input cutoff frequency.

92 ~~93~~. The television receiver as set forth in Claim 91 ~~92~~ wherein said minimum frequency value of said first reference signal is greater than said input cutoff frequency.

93 ~~94~~. The television receiver as set forth in Claim 88 ~~89~~ wherein said second mixer mixes said second reference signal and a first filtered signal on said output of said first bandpass filter to thereby generate a second IF signal.

94 ~~95~~. The television receiver as set forth in Claim 93 ~~94~~ wherein said second operating frequency is fixed.

95 ~~96~~. The television receiver as set forth in Claim 94 ~~95~~ wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

96 ~~97~~. The television receiver as set forth in Claim 94 ~~95~~ wherein said minimum frequency value of said first reference signal is greater than an input cutoff frequency of said RF signal.

Sub a11

98. A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value;

a first bandpass filter coupled to an output of said first mixer; and

a second mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

~~99~~ 98. The television receiver as set forth in Claim ~~98~~ 97 wherein said minimum frequency value is greater than an input cutoff frequency of an RF signal outputted ~~received~~ from said RF signal source.

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~~100~~ 99. The television receiver as set forth in Claim ~~99~~ 98 wherein said first mixer subtractively mixes said RF signal and said first reference signal to thereby produce a first IF signal, wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal and said first predetermined frequency is greater than said selected carrier frequency.

~~101~~ 100. The television receiver as set forth in Claim ~~100~~ 99 wherein said selected carrier frequency appears at

said first predetermined frequency in an output passband signal of said bandpass filter.

~~102~~ 101. The television receiver as set forth in Claim ~~101~~ 100 wherein said second mixer mixes said output passband signal and said second reference signal to thereby produce a second IF signal, wherein said selected carrier frequency appears in said second IF signal at a second predetermined frequency and said second predetermined frequency is less than said first predetermined frequency and said second operating frequency.

~~103.~~ 102. A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a first bandpass filter to produce an output passband signal; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

~~104.~~ 103. The method as set forth in Claim ~~103~~ 102 including the further step of filtering out of the RF signal all frequencv components above an input cutoff frequency.

~~105.~~ 104. The method as set forth in Claim ~~104~~ 103 wherein the first operating frequency is greater than the input cutoff frequency.

~~106.~~ 105. The method as set forth in Claim ~~105~~ 104 wherein the step of mixing the RF signal with the first reference signal is a subtractive mixing of the RF signal and the first reference signal.

~~107.~~ 106. The method as set forth in Claim ~~106~~ 105 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a first predetermined frequency and wherein the first predetermined frequency is higher than the ~~RF signal~~ selected carrier frequency.

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107 ~~108.~~ The method as set forth in Claim 106 ~~107~~ wherein the first predetermined frequency appears in the output passband signal of the first bandpass filter.

Sub a12

109. A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a bandpass filter to produce an output passband signal; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

109 ~~110~~. The method as set forth in Claim 108 ~~109~~ wherein the first operating frequency is greater than the input cutoff frequency.

SUB B9

~~111~~. The method as set forth in Claim 110 wherein the step of mixing the RF signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in the RF signal appears in the first IF signal at a first predetermined frequency and wherein said first predetermined frequency is higher than the RF signal.

111 ~~112~~. The method as set forth in Claim 110 ~~111~~ wherein the selected carrier signal appears in the output passband signal at the first predetermined frequency.

112 ~~113~~. The method as set forth in Claim 108 ~~109~~ wherein the step of mixing the output passband signal with the

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second reference signal is a down conversion mixing, wherein ~~the~~ a selected carrier signal appearing at a ~~the~~ first predetermined frequency in the output passband signal appears in the second IF signal at a second predetermined frequency and wherein said second predetermined frequency is lower than said first predetermined frequency.

~~114~~ 113. The method as set forth in Claim ~~109~~ 108 including the further step of filtering the second IF signal with a second bandpass filter.

~~115~~ 114. The method as set forth in Claim ~~114~~ 113 wherein the second bandpass filter has a passband between 15 Megahertz and 21 Megahertz.

~~116~~ 115. The method as set forth in Claim ~~115~~ 114 wherein the second bandpass filter is a Nyquist slope filter having a cutoff frequency of 20 Megahertz.

add a13

add D1

2836-P223US

PATENT
Sole Inventor
Original

DECLARATION AND POWER OF ATTORNEY

As a below named inventor, I declare that:

My residence, post office address and citizenship are as stated below next to my name, that I believe I am the original, first and sole inventor of the subject matter which is claimed and for which a patent is sought on the invention or design entitled "MONOLITHIC TELEVISION TUNER" the specification of which is attached hereto; that I have reviewed and understand the contents of the above-identified specification, including the claims, as amended by any amendment referred to above; and that I acknowledge my duty to disclose information known to be material to the patentability of a pending claim in this application in accordance with 37 C.F.R. 1.56(a).

I hereby claim the benefit under Title 35 United States Code § 120 of any United States application(s) listed below and, insofar as any subject matter of any claim of this application is not disclosed in the prior United States Application, I acknowledge the duty to disclose material information as defined in Title 37, Code of Federal Regulations § 1.56(a) which occurred between the filing date of the prior application and the national or PCT international filing date of this application: N/A

Serial No.	Filing Date

We hereby appoint:

David H. Tannenbaum	Reg. No. 24,745
N. Elton Dry	Reg. No. 22,600
George E. Clark	Reg. No. 25,133
Kelly K. Kordzik	Reg. No. 36,571
James J. Murphy	Reg. No. 34,503
Elizabeth R. Hall	Reg. No. 37,344
Philip T. Golden	Reg. No. 37,631

7

of the firm of Winstead Sechrest & Minick P.C., my attorneys with full power of substitution and revocation, to prosecute this application and to transact all business in the United States Patent and Trademark Office connected therewith, and to file and prosecute any international patent applications filed thereon before any international authorities under the Patent Cooperation Treaty.

Send Correspondence To:	Direct Telephone Calls To
Winstead Sechrest & Minick P.C. 5400 Renaissance Tower 1201 Elm Street Dallas, Texas 75270	David H. Tannenbaum at (214) 745-5354 Atty. Docket No. 2836-P223US

I declare that all statements made herein of my own knowledge are true and that all statements made on information and belief are believed to be true; and further that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under Section 1001 of Title 18 of the United States Code, and that such willful false statements may jeopardize the validity of the application or any patent issuing thereon.

1-80

Full name of the sole inventor

Robert Rudolf Rotzoll

Inventor's signature	Robert Rudolf Rotzoll
Date	April 13, 1995
Residence (City, County, State)	Allen, Collin County, Texas
Citizenship	U.S.A. TX
Post Office Address	201 Fairfax Drive, Allen, Texas 75002

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238:2836-P223US

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PATENT

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APR 21 1995
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APPLICATION FOR U.S. PATENT
TRANSMITTAL FORM

THE COMMISSIONER OF PATENTS
AND TRADEMARKS
Washington, D.C. 20231

Docket No. 2836-P223US

Sir:

Transmitted herewith for filing is the patent application of:

Inventor(s): Robert Rudolf Rotzoll

For: MONOLITHIC TELEVISION TUNER

Enclosed are: 4 Sheets of Informal Drawings

An assignment of the invention to:
Cirrus Logic, Inc.

FEE CALCULATION					FEE
	Number		Number Extra	Rate	Basic Fee $ 730.00
Total Claims	116	- 20 =	96	x $22 =	$ 2,112.00
Independent Claims:	9	- 3 =	6	x $76 =	$ 456.00
Multiple Dependent Claims:					$
				Total Filing Fee =	$ 3,298.00

Enclosed is a check in the amount of $3,298.00. Please charge any additional fees or credit any overpayment to Deposit Account No. 23-2426 of Winstead Sechrest & Minick P.C. A duplicate copy of this sheet is enclosed.

Please return the original Assignment document to the undersigned attorney for Applicant following recordal of same and address it to: WINSTEAD SECHREST & MINICK P.C., 5400 Renaissance Tower, 1201 Elm Street, Dallas, Texas 75270.

April 21, 1995
Date

David H. Tannenbaum
Attorney for Applicant
Registration No. 24,745

041995tlml
233:2836-P223US

[redacted]

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PATENT

MAIL ROOM
2836-P223US
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APR 21 1995
PAT. & TRADEMARK OFF.

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IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application of:	Robert Rudolf Rotzoll
Filed:	April 21, 1995
For:	MONOLITHIC TELEVISION TUNER

HONORABLE COMMISSIONER OF
PATENTS AND TRADEMARKS
Washington, D.C. 20231

Dear Sir:

CERTIFICATE OF MAILING BY EXPRESS MAIL

I certify that this Application, Informal Drawings, Declaration and Power of Attorney, and Assignment documents are being deposited with the United States Postal Service "Express Mail Post Office to Addressee" service under 37 C.F.R. 1.10 on April 21, 1995, as addressed to the Commissioner of Patents and Trademarks, Washington, D.C. 20231.

Terri Massengale

Terri Massengale

Express Mail Receipt No.

TB654431162US

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233:2836-P223US

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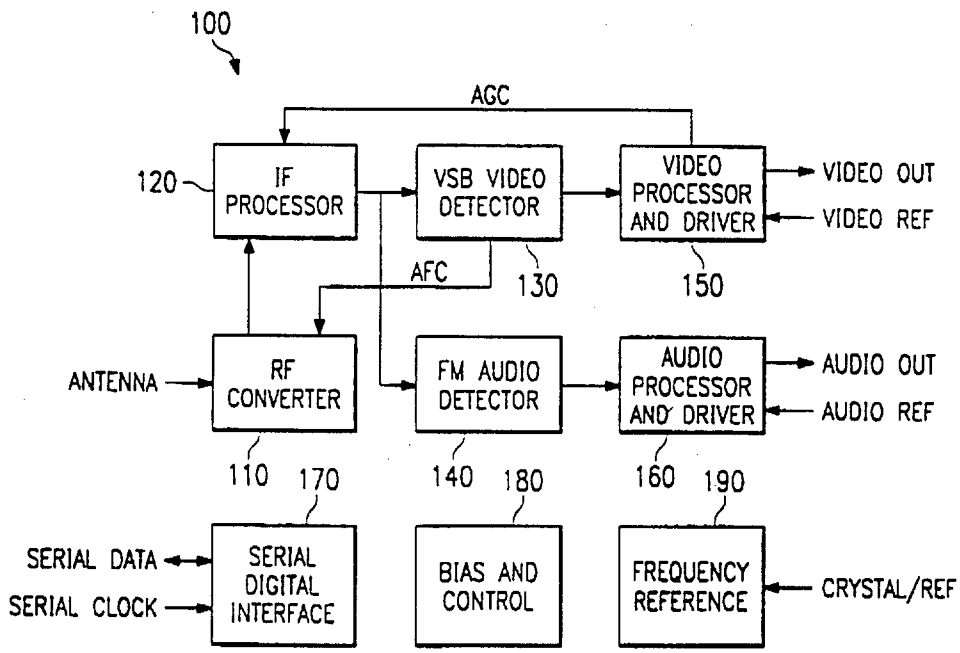

FIG. 1

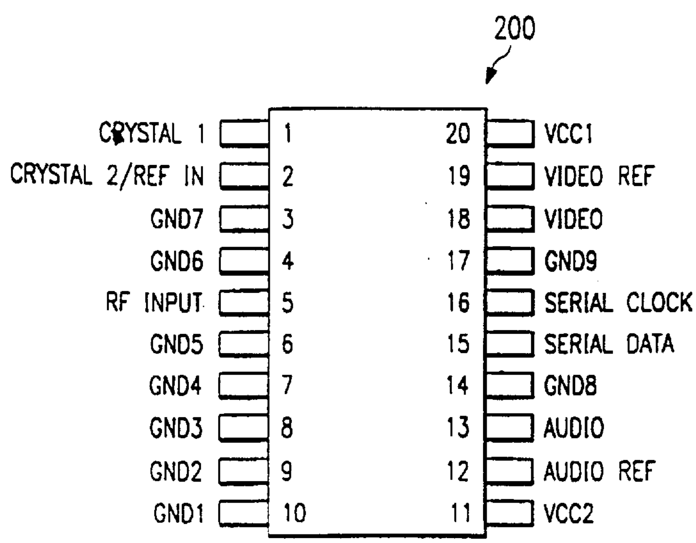

FIG. 2

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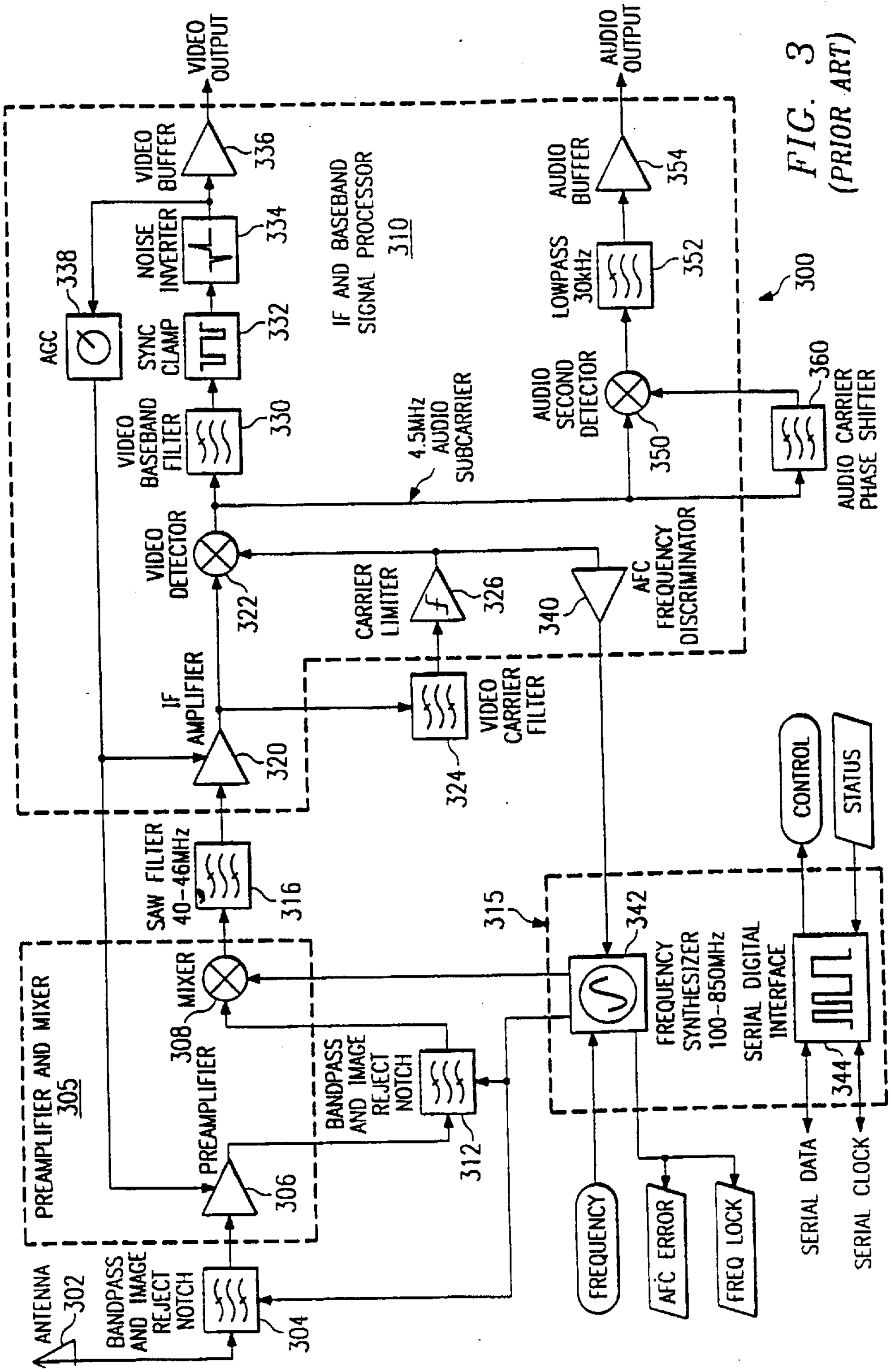

FIG. 3
(PRIOR ART)

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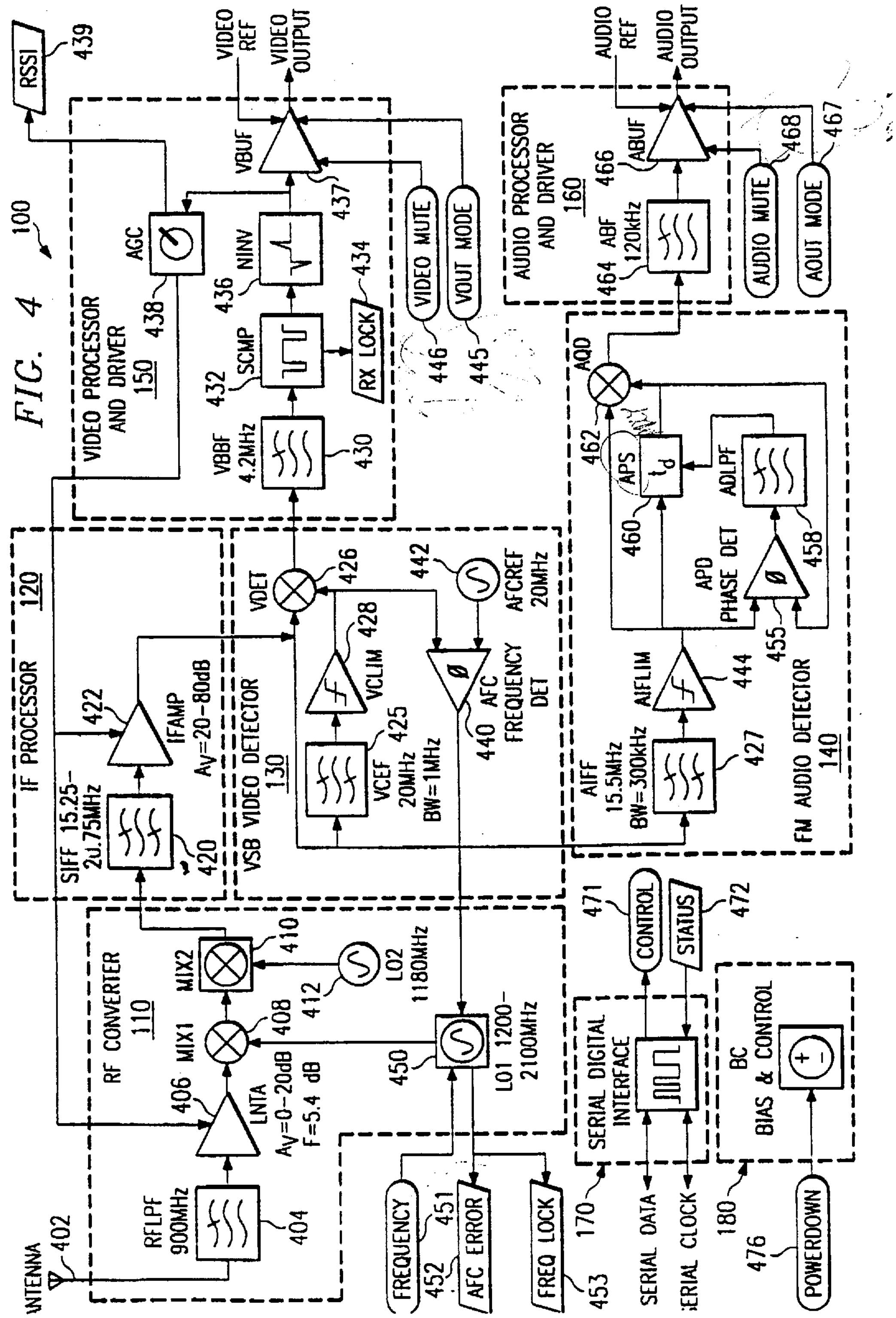

FIG. 4

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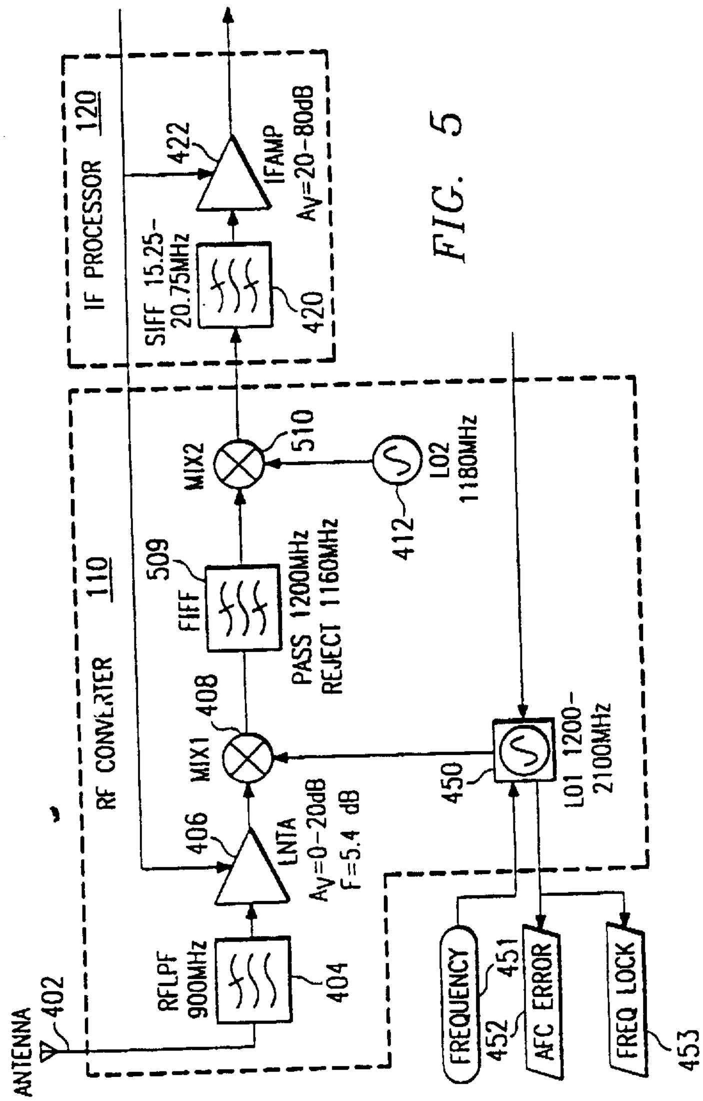

FIG. 5

Notes:

TUNE 048246

TAB 3

Office Action Summary	Application No. 08/426,080	Applicant(s) Rotzoll
	Examiner Brian Johnson	Group Art Unit 2602

☐ Responsive to communication(s) filed on ______.

☐ This action is **FINAL**.

☐ Since this application is in condition for allowance except for formal matters, prosecution as to the merits is closed in accordance with the practice under *Ex parte Quayle*, 1935 C.D. 11; 453 O.G. 213.

A shortened statutory period for response to this action is set to expire 3 month(s), or thirty days, whichever is longer, from the mailing date of this communication. Failure to respond within the period for response will cause the application to become abandoned. (35 U.S.C. § 133). Extensions of time may be obtained under the provisions of 37 CFR 1.136(a).

Disposition of Claims

☒ Claim(s) *1-116* is/are pending in the application.

Of the above, claim(s) ______ is/are withdrawn from consideration.

☐ Claim(s) ______ is/are allowed.

☒ Claim(s) *1-116* is/are rejected.

☐ Claim(s) ______ is/are objected to.

☐ Claims ______ are subject to restriction or election requirement.

Application Papers

☒ See the attached Notice of ~~Draftsperson's Patent Drawing Review, PTO-948.~~ Informal Drawings, Substitute PTO-948

☐ The drawing(s) filed on ______ is/are objected to by the Examiner.

☐ The proposed drawing correction, filed on ______ is ☐ approved ☐ disapproved.

☐ The specification is objected to by the Examiner.

☐ The oath or declaration is objected to by the Examiner.

Priority under 35 U.S.C. § 119

☐ Acknowledgement is made of a claim for foreign priority under 35 U.S.C. § 119(a)-(d).

☐ All ☐ Some* ☐ None of the CERTIFIED copies of the priority documents have been

☐ received.

☐ received in Application No. (Series Code/Serial Number) ______.

☐ received in this national stage application from the International Bureau (PCT Rule 17.2(a)).

*Certified copies not received: ______.

☐ Acknowledgement is made of a claim for domestic priority under 35 U.S.C. § 119(e).

Attachment(s)

☒ Notice of References Cited, PTO-892

☒ Information Disclosure Statement(s), PTO-1449, Paper No(s). 3

☐ Interview Summary, PTO-413

☒ Notice of ~~Draftsperson's Patent Drawing Review, PTO-948~~ Informal Drawings, Sub. PTO-948

☐ Notice of Informal Patent Application, PTO-152

— SEE OFFICE ACTION ON THE FOLLOWING PAGES —

1. The following is a quotation of the appropriate paragraphs of 35 U.S.C. § 102 that form the basis for the rejections under this section made in this Office action:

> A person shall be entitled to a patent unless --
> (b) the invention was patented or described in a printed publication in this or a foreign country or in public use or on sale in this country, more than one year prior to the date of application for patent in the United States.

Claims 1, 2, 4-7, 9-13, 17, 18, 24-27, 29-33, 35-37, 39, 40, 43, 46, 47, 84-89, and 91-113 are rejected under 35 U.S.C. Section 102(b) as being anticipated by Ash. Considering claim 1, Ash (Fig. 1, Ref. 10) discloses a receiver input coupled to a receiver source, a first reference signal (Fig. 1, Ref. 23), a first mixer having a first input coupled to the receiver input and a second input coupled to the first reference signal (Fig. 1, Ref. 20), a second reference signal (Fig. 1, Ref. 34), and a second mixer (Fig. 1, Ref. 32) connected as claimed by the applicant.

As for claim 24, Ash (Fig. 1, Ref. 10) discloses a receiver input coupled to an RF signal source, a first mixer having a first input coupled to the receiver input (Ash Fig. 1, Ref. 20) and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value (Ash, Fig. 1, Ref. 22), a second mixer having a first input coupled to an output of the first mixer and a second

input coupled to a second reference signal having a second operating frequency (Ash Fig. 1, Ref. 32), wherein the second operating frequency is fixed (Ash, col. 6, lines 24-26).

Considering claim 29, Ash (Fig. 2b) discloses mixing the RF signal with a first reference signal having a first operating frequency to produce a first IF signal and mixing the first IF signal with a second reference signal having a second operating frequency to produce a second IF signal (Ash, Fig. 1, Ref. 32).

Regarding claim 35, Ash (Fig. 1, Ref. 12) discloses filtering all frequency components above an inp t cutoff frequency out of the received RF signal to thereby produce a first filtered signal, amplifying the first filtered signal (Fig. 1, Ref. 17), subtractively mixing the signals to produce a first IF signal (Fig. 2b), and mixing the first IF signal with a second reference signal to produce a second IF signal (Fig. 1, Ref. 32).

Considering claim 43, Ash discloses an input filter coupled to an RF signal source (Fig. 1, Ref. 12), a first amplifier (Fig. 1, Ref. 17), a first mixer (Fig. 1, Ref. 20), a first oscillator (Fig. 1, Ref. 22), a second mixer (Fig. 1, Ref. 32), a second oscillator (Fig. 1, Ref. 35), and a second filter (Fig. 1, Ref. 71).

Claim 84 is rejected for the reasons set forth regarding claim 1. Additionally, Ash (Fig. 1, Ref. 28) discloses a first bandpass filter coupled to an output of the first mixer.

As for claim 98, Ash discloses a receiver input coupled to an RF signal source (Fig. 1, Ref. 10), a first mixer coupled to the receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein the first operating frequency is variable between a minimum frequency and a maximum frequency value (Fig. 1, Refs. 20-23), a bandpass filter (Fig. 1, Ref. 28), and a second mixer having a first input coupled to an output of the bandpass filter and a second input coupled to the second reference signal (Fig. 1, Ref. 32), wherein the second operating frequency is fixed (Ash col. 6, lines 24-26).

As for claim 103, Ash discloses the method of mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal (Fig. 1, Ref. 20), filtering the first IF signal (Fig. 1, Ref. 28), and mixing the output passband with a second reference signal to produce a second IF signal.

Claim 109 is rejected for the reasons set forth concerning claims 1 and 84. Additionally, Ash (Fig. 1, Ref. 17) also

discloses the amplification of the first signal as claimed by the applicant.

As for claims 2, 6, 32, 85, 88 and 106, Ash (Fig. 2b) discloses the claimed subtractive mixer.

Regarding claims 4, 11, 25, 31, 36, 86, 93, 97, 99, 105, and 110, Ash (Fig. 1, Ref. 22 and col. 1, lines 18-29) discloses the receiver wherein a minimum frequency value of the first reference signal is greater than an input cutoff frequency of an RF signal received from the RF signal source.

Concerning claims 5 and 87, Ash (Fig. 1, Ref. 22) discloses a television receiver wherein the first operating frequency is variable between a minimum frequency value and a maximum frequency value.

As for claims 7, 37, 89, 107, and 111, Ash (Fig. 2b) discloses the selected carrier frequency in the RF signal appearing at a first predetermined frequency in the first IF signal.

Regarding claims 9 and 91, Ash (col. 5, lines 46-49) discloses the selection of the carrier frequency by varying the first operating frequency.

As for claims 10, 30, 92, and 104, Ash (Fig. 1, Refs. 12, 13, and 60) discloses the first mixer coupled to the receiver input by an input filter.

Concerning claims 12 and 17, Ash (Fig. 1, Ref. 32) shows the claimed second mixer which generates the second IF signal.

As for claims 13, 18, and 95, Ash shows a fixed operating frequency (col. 6, lines 24-26).

Regarding claim 96, Ash (col. 6, line 30) discloses the selected carrier frequency in the RF signal appearing at a second predetermined frequency in the second IF signal.

As for claim 33, Ash (col. 5, line 50) discloses the step of mixing the RF signal with the first reference signal in an up-conversion process, wherein a selected carrier frequency in the RF signal appears in the first IF signal at a frequency higher than the RF signal.

Regarding claim 40, Ash (col. 8, line 35) discloses the step of filtering the second IF signal with a bandpass filter.

As for claim 46, Ash (col. 7, line 27) discloses the gain of the first amplifier automatically controlled by an automatic gain control circuit of the television receiver.

Regarding claims 39, 101, 108, 112, and 113, Ash (Fig. 2b) illustrates the second mixer mixing the output passband signal and the second reference signal to produce a second IF signal, wherein the selected carrier frequency appears in the second IF signal at a predetermined frequency less than the first predetermined frequency and the second operating frequency.

Claim 26 is rejected for the reasons set forth concerning claims 6 and 7 above.

Claim 27 is rejected for the reasons set forth concerning claims 17 and 20 above.

Claim 47 is rejected for the reasons set forth regarding claims 29, 32, and 33 above.

Claim 94 is rejected for the reasons set forth concerning claims 12 and 84 above.

Claim 100 is rejected for the reasons set forth concerning claims 6, 7, and 26.

Claim 102 is rejected for the reasons set forth concerning claims 12, 15, and 84.

2. The following is a quotation of 35 U.S.C. § 103 which forms the basis for all obviousness rejections set forth in this Office action:

> A patent may not be obtained though the invention is not identically disclosed or described as set forth in section 102 of this title, if the differences between the subject matter sought to be patented and the prior art are such that the subject matter as a whole would have been obvious at the time the invention was made to a person having ordinary skill in the art to which said subject matter pertains. Patentability shall not be negatived by the manner in which the invention was made.
>
> Subject matter developed by another person, which qualifies as prior art only under subsection (f) or (g) of section 102 of this title, shall not preclude patentability under this section where the subject matter and the claimed invention were, at the time the invention was made, owned by the same person or subject to an obligation of assignment to the same person.

Claims 8, 41, 42, 44, 45, 48, 50-76, 90 and 114-116 are rejected under 35 U.S.C. § 103 as being unpatentable over Ash. Regarding claims 8 and 90, Ash (Fig. 1, Ref. 68) allows for the use of higher frequency ranges in the neighborhood of 1200 MHz. It would have been obvious to one skilled in the art at the time the invention was made to use higher frequencies in order to allow the receiver to function within a broader range and allow the user to tune into a wider variety of broadcast signals.

As for claims 41, 42, and 51-54 the use of various filtering techniques at different frequencies is commonly known in the art of signal processing and would have been obvious to one skilled in the art at the time the invention was made in order to achieve the desired result.

Considering claim 44, Ash (col. 7, line 33) discloses a low pass filter with a -3db cutoff frequency of 936 MHz. Although the Applicant claims a cutoff frequency of 900 MHz, the Ash design is within a range of appropriate frequencies that would have been obvious to one skilled in the art at the time the invention was made.

Regarding claims 45 and 56, various amplification devices with appropriate noise figures would have been readily obvious to one skilled in the art at the time the invention was made.

Concerning claims 48 and 50, various devices with appropriate frequency responses could be used for the oscillators and would have been known to one skilled in the art at the time the invention was made.

Considering claim 55, Ash (col. 6, line 42) utilizes the synchronization pulses to ensure that the signal processing steps occur in a compatible fashion. It would have been obvious to one

skilled in the art at the time the invention was made to use the sync pulses in this manner because the sync pulses are commonly used in television signal processing to control timing of various electronic components in a television circuit.

Regarding claim 57, Ash (col. 15, lines 24-25) discloses the gain of the second amplifier automatically controlled by the automatic gain control circuit. Automatic gain control is commonly used in signal processing and would have been obvious to one skilled in the art at the time the invention was made.

As for claim 58, Ash (col. 8, lines 28-41) discloses a filtering and mixing process by which the video baseband signal is extracted. It would have been obvious to one skilled in the art to utilize a bandpass filter (Fig. 1, Ref. 38) along with a mixer (Fig. 1, Ref. 42) to obtain the video baseband signal as claimed in order to allow for separate processing of the video and audio components of the television signal. This type of signal detection is common within the art of television signal processing and would have been readily obvious to one skilled in the art at the time the invention was made.

Considering claim 59, Ash (Fig. 1, Ref. 45) discloses a means of providing a fixed amplitude reference carrier signal to the mixer used for producing the video baseband signal. The

reference signal would have been an inherent feature in the synchronous detector design used to reduce the selected channel video signals to baseband. Therefore, the use of a reference carrier would have been obvious to one skilled in the art at the time the invention was made.

Regarding claim 60, Ash (col. 6, lines 41-57) describes the use of frequency detection circuitry. This circuitry would have been obvious to one skilled in the art at the time the invention was made because frequency detection is a commonly used technique for isolating certain portions of a composite signal for further processing. In this fashion, one could easily perform necessary adjustments to video and audio data separately and provide better images and sounds to a user.

Concerning claim 61, the use of a feedback control system is common within the art of signal processing and would have been readily obvious to one skilled in the art at the time the invention was made. Such feedback provides better reliability and reduces noise throughout the circuitry.

As for claim 62, Ash (Fig. 1, Ref. 71) discloses the low pass filter coupled to the third mixer which cuts off all signal components higher in frequency than the video baseband signal. Such filtering is common in television signal processing and

would have been obvious to one skilled in the art at the time the invention was made.

Regarding claims 63-66, filtering noise and ensuring that the desired video signal does not drop below a certain level are common methods of processing a video signal before displaying the signal and would have been obvious to one skilled in the art. In this manner, the viewer will not be subject to inconsistent data causing distorted images on the screen.

Considering claims 67 and 68, Ash (col. 15, lines 24-25) discloses the claimed automatic gain control circuit, which would have been obvious to a skilled artisan due to the prevalence of automatic gain control circuits in television signal processing. Ash (col. 5, lines 36-40) applies the gain control circuit to the first amplifier. Such gain control could obviously be applied to any amplifier within the system.

Claims 69-74 are rejected as being obvious methods of enhancing the video signal before displaying the data. A variety of filtering and amplification techniques are commonly used in the art of television signal processing in order to ensure that the data is displayed with as little noise as possible and that the data is timed so that images are displayed without distortion.

Regarding claims 75 and 76, Ash (Fig. 20, Ref. 46) discloses a filter for extracting the audio component from the video output signal. Although Ash does not rely upon the frequencies as claimed in claim 76, it would have been obvious to one skilled in the art at the time the invention was made use any appropriate frequency for design purposes.

Claims 114-116 are rejected as being commonly used noise reduction filters that would have been obvious to one skilled in the art at the time the invention was made.

3. Claims 3, 14-16, 19-23, 28, 34, 38, and 49 are rejected under 35 U.S.C. § 103 as being unpatentable over Ash in view of Ma. Regarding claims 3, 14, 19, 28, 34, 38, and 49, Ma (col. 4, lines 44-47) discloses a second mixer which is an image rejection mixer. It would have been obvious to one skilled in the art at the time the invention was made to use a subtractive mixer in order to provide more accurate tuning in the receiver design.

As for claims 15 and 20, Ash (col. 6, line 30) discloses the selected carrier frequency in the RF signal appearing at a second predetermined frequency in the second IF signal.

Concerning claim 16, Ash (Fig. 1, Ref. 22 and col. 1, lines 18-29) discloses a minimum frequency value of the first reference

signal as being greater than the input cutoff frequency of the RF signal received from the RF signal source.

Regarding claim 21, Ash (Fig. 2b) discloses the claimed second predetermined frequency as the difference between the first predetermined frequency and the second operating frequency.

Considering claim 22, Ash illustrates a television receiver wherein the second operating frequency is less than the first predetermined frequency and the frequency of the image signal rejected by the second mixer is less than the second operating frequency by an amount equal to the second predetermined frequency (Ash, Fig. 2d).

As for claim 23, Ash (Ref. 68) illustrates that a range of operating frequencies are available for use with the Ash design, and the frequencies claimed by the applicant would have been obvious to one skilled in the art at the time the invention was made.

4. Claims 77-81 are rejected under 35 U.S.C. § 103 as being unpatentable over Ash in view of Nordholt. As for claim 77, Nordholt discloses the audio phase detector circuitry (Fig. 1, PLL2) and the mixer as claimed by the applicant for producing the appropriate audio signal response (Fig. 1, SD). It would

have been obvious to one skilled in the art at the time the invention was made to isolate the audio signal in order to process the audio signal separately from the video signal.

Considering claim 78, Nordholt (col. 4, lines 34-49) illustrates a loop operating in phase quadrature. Such a design would have been obvious to one skilled in the art at the time the invention was made in order to provide for an efficient method of detecting the audio signal to be ultimately heard by the user.

Claims 79-81 are rejected as being obvious filtering and amplification techniques used to provide a noise free signal at the required power level, which would have been readily obvious to one skilled in the art at the time the invention was made.

5. Claims 82 and 83 are rejected under 35 U.S.C. § 103 as being unpatentable over Ash in view of Nordholt as applied to claim 80 above, and further in view of Sussman. Regarding claim 82, Sussman (Fig. 1, Ref. 28) discloses the claimed muting means which would have been obvious to one skilled in the art at the time the invention was made because muting means are currently available on most TV sets in order to provide a quick method of turning the volume to zero by pressing one button.

Considering claim 83, Sussman (Fig. 1, Ref. 27) provides a means of varying the volume of the television. Volume control is inherent in most television designs and would have been readily obvious to one skilled in the art at the time the invention was made.

6. Any inquiry concerning this communication or earlier communications from the examiner should be directed to Brian Johnson whose telephone number is (703) 308-7583. The examiner can normally be reached on Monday-Friday from 9am to 4pm.

If attempts to reach the examiner by telephone are unsuccessful, the examiner's supervisor, James Groody, can be reached on (703) 305-4702. The fax phone number for this Group is (703) 305-9509.

Any inquiry of a general nature or relating to the status of this application or proceeding should be directed to the Group receptionist whose telephone number is (703) 305-4700.

VICTOR R. KOSTAK
PRIMARY EXAMINER
ART UNIT 262

Attachment Paper No. 4

The drawings submitted with this application were declared informal by the applicant. Accordingly they have not been reviewed by a draftsperson at this time. When formal drawings are submitted, the draftsperson will perform a review.

Direct any inquires concerning drawing review to the Drawing Review Branch (703) 305-8404.

Notes:

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TAB 4

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47681-P023US-954919

PATENT

#5/a
mail
6-4-96

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application:	Robert Rudolf Rotzoll
Serial No.:	08/426,080
Filed:	April 21, 1995
Art Unit:	2602
Examiner:	Johnson
For:	MONOLITHIC TELEVISION TUNER

Andrade
6-4-96

Assistant Commissioner
for Patents
Washington, D.C. 20231

I hereby certify that this correspondence is, on the date shown below, being deposited with the United States Postal Service with sufficient postage as first class mail, in an envelope addressed to the Assistant Commissioner for Patents, Washington, DC 20231.

Date deposited: May 8, 1996

Signature: Elva Wilson

Printed Name: Elva Wilson

AMENDMENT

Dear Sir:

In response to the Office Action, mailed February 9, 1996, Applicant respectfully requests that the Examiner reconsider the application in view of the amendments and comments set forth below.

In the Specification

On page 11 please change the description of FIGURES 3 and 4 to read as follows:

FIGURE 3 is a detailed block diagram of a state-of-the-art television tuner found in the prior art; [and]

FIGURE 4 is a detailed block diagram of the circuit components of the present invention[.];

330 SB 05/16/96 08426080
1 102 78.00 CK
1 103 66.00 CK

Please add a description for FIGURE 5 and for a new FIGURE 6 as follows:

a2

FIGURE 5 shows an alternate embodiment of the RF front-end of the present invention; and

FIGURE 6 is a detailed block diagram of MIX2 of FIGURE 4.

On page 28, lines 22-29 should be changed to read as follows:

a3

--...cell multipliers 610 and 612 at MIX2 input and using fairly high-order balanced RC-phase shifters 611, 613, 614 with very low component sensitivity with resulting phase errors via summer 615 of less than 0.2...--

In the Claims

Please amend the claims as shown.

In claim 1, page 48, line 12, after the word "coupled" please insert the word --directly--.

a4

11 ~~12~~. (Amended) The television receiver as set forth in Claim 6 ~~7~~ wherein said second mixer mixes said second reference signal and said first IF signal to thereby generate a second IF signal[.] containing all of the channels of information as contained in said ~~the original~~ RF signal.

a5

17. (Amended) The television receiver as set forth in Claim 1 further comprising a first IF signal generated by said first mixer wherein said second mixer mixes said second reference signal and said first IF signal to generate a second IF signal[.] containing all of the channels of information as contained in the original RF signal.

SUB C3

24. (Amended) A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer physically located on an integrated circuit substrate having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value; and

a6

a second mixer physically located on the same integrated circuit substrate as the first mixer and having a first input coupled without leaving the substrate to an output of said first mixer and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

SUB B4

29. (Amended) A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

a7

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal[.], said IF mixing occurring prior to eliminating any channels from the signal.

SUB B5

35. (Amended) A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

a8

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal[.], said IF mixing occurring prior to eliminating any channels from the signal.

sub C5

43. (Amended) A television receiver comprising:

an input filter coupled to an RF signal source;

a first amplifier coupled to an output of said input filter;

a first mixer having a first input and a second input, wherein said first input of said first mixer is coupled to an output of said first amplifier;

a9

a first oscillator coupled to said second input of said first mixer, wherein said first oscillator generates a first reference signal having a first operating frequency;

a second mixer having a first input and a second input, wherein said first input of said second mixer is coupled to an output of said first mixer without the connection of a saw filter therebetween;

a second oscillator coupled to said second input of said second mixer, wherein said second-oscillator generates a second reference signal having a second operating frequency; and

a second filter coupled to an output of said second mixer.

sub C9

84. (Amended) A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a10

a first bandpass filter coupled to an output of said first mixer; the bandpass filter passing more than one channel and not forming an image rejection for a subsequent mixer;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to said second reference signal.

97 ~~98~~. (Amended) A television receiver comprising:

a receiver input coupled to an RF signal source;

a first mixer having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value;

a first bandpass filter constructed in the same substrate as said first mixer and coupled to an output of said first mixer; and

a11 a second mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

1? ~~103~~. (Amended) A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a first bandpass filter to produce an output passband signal having at least four channels; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

108 ~~109~~. (Amended) A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

a12 subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal;

filtering the first IF signal in a bandpass filter to produce an output passband signal having at least four channels; and

mixing the output passband signal with a second reference signal having a second operating frequency to thereby produce a second IF signal.

Please add new claims 117, 118 and 119 as follows:

AB10

--117. (New) The method of extracting a single communications channel from a large number of such channels wherein each such channel has a defined frequency range, and wherein the large number of channels are communicated within an upper and a lower frequency boundary, the method comprising the steps of:

from an input signal containing all of the large number of channels removing all frequencies greater than the upper frequency boundary;

translating all of the channels contained within the input signal to a translated frequency range such that a desired single channel to be extracted from the input signal is positioned centrally within the translated frequency range, the translated frequency range being positioned higher in the frequency spectrum than the upper frequency boundary;

removing from the translated input signal all channels extending from the center of the translated frequency range four channels or more in each frequency direction;

113

translating the remaining channels within the input signal to a second translated frequency range, such that the desired single channel to be extracted from the input signal is positioned centrally within the second translated frequency range having a low end frequency boundary defined as high enough not to cause phase distortion in any subsequent filtering and a high end frequency boundary defined as the frequency beyond which a filter with proper bandwidth, noise and distortion can be integrated in a substrate; and

removal of all remaining frequencies which are not part of the desired single channel.--

--~~118~~ 117. (New) The method set forth in claim ~~117~~ 116 wherein the last-mentioned removal step includes a receiver attenuation characteristic such that the relative amplitude of low and high frequencies within the desired single channel are corrected for vestigial sideband demodulation.--

--~~119~~ 118. (New) The method set forth in claim ~~117~~ 116 wherein the low end and high end boundaries of the second translated frequency range are defined by the implementation of an integrated circuit continuous time filter capable of achieving single channel TV signal extraction and receiver attenuation characteristics.--

Remarks

Upon reviewing the specification, it became clear that it would be helpful if the details of circuit MIX2 410 were to be shown in a figure in addition to the explanation given in the specification. Accordingly, FIGURE 6 has been added. Support for this figure is shown at pages 28, et seq., which pages have been amended to add the element numbers shown in new FIGURE 6. No new matter has been added by this figure. In addition, the description of FIGURE 5 was added to the "Summary" on page 11, as it was originally omitted. Support for this change is shown at page 23, lines 4, et seq.

Applicant understands that there are a large number of claims in this application and appreciates the Examiner's thoroughness in responding to all of the claims.

Claims 1, 2, 4-7, 9-13, 17, 18, 24-27, 29-33, 35-37, 39, 40, 43, 46, 47, 84-89 and 91-113 were rejected under 35 U.S.C. § 102(b) as being anticipated by Ash. With respect to claim 1, the Examiner has said:

> "Ash (Fig. 1, Ref. 10) discloses a receiver input coupled to a receiver source, a first reference signal (Fig. 1, Ref. 23), a first mixer having a first input coupled to the receiver input and a second input coupled to the first reference signal (Fig. 1, Ref. 20), a second reference signal (Fig. 1, Ref. 34), and a second mixer (Fig. 1, Ref. 32) connected as claimed by the applicant."

While at first glance, the Examiner's comments seem to be accurate, a further reading of claim 1 in conjunction with the specification reveals a significant difference between the Ash teaching and the claimed invention. These differences stem primarily from the fact that the claimed invention was designed to be implemented on a single substrate, therefore resulting in certain demands of frequency separation and detection some of which heretofore have been accomplished by devices <u>external</u> to the substrate. This has dictated certain design concept changes from prior systems such as Ash.

For example, in Ash, as the Examiner pointed out, it is MIXER32 which is the equivalent of the second MIXER as set forth in claim 1 of the instant invention. However, claim 1 recites that the second MIXER has a first input coupled to an

output of the first MIXER. The intention of this recitation in claim 1 was that the output of MIXER1 would connect directly to the input of MIXER2 without an external filter. Claim 1 has been modified to include the word "directly" to avoid any misunderstanding as to the nature of the connection.

As shown in FIGURE 1 of Ash, several elements, such as switch 25, and more importantly, filter 28 are interposed between MIXER1 (element 20) and second MIXER (element 32). In Ash filter 28 is a channel selection filter and is used to select out all of the undesired channels before the second MIXER. As noted by the dotted lines of FIGURE 1 of Ash, filter 28 is a filter which is not (and cannot be) constructed on the same substrate as the mixers. A filter of a type used by Ash to eliminate all unwanted channels is not practical to construct on the same substrate as the mixers. In the instant invention where a single substrate is utilized, a different circuit construction must be designed. One design change requires both mixers to be connected without first removing all of the unwanted channels. None of the prior art shows such a circuit, and accordingly, claim 1 should be held allowable.

Claim 24 has been amended to specifically recite that two mixers are located on the same substrate. As discussed above, Ash does not show such a circuit nor can the Ash circuit function in this manner. Accordingly, claim 24 should be held allowable, as amended.

Claim 29 has been amended to recite that the second mixing step occurs prior to eliminating any channels from the signal. Ash, as discussed above, removes the unwanted channels from the signal prior to the second mixing step. Accordingly, claim 29 should be held allowable, as amended.

Claim 35 has been amended in the same manner as claim 29 to clarify that the second stage mixing occurs prior to the elimination of any channels from the IF signal. Ash does not teach such a concept. Accordingly, claim 35 should be held allowable, as amended.

Claim 43 has been amended to specifically recite that the connection between the first mixer and the second mixer does not have a saw filter connected therebetween. As discussed above, this is unique to the claimed invention and not shown in the prior art. Accordingly, claim 43 should be held allowable, as amended.

Claim 84 was rejected for the same reasons as claim 1. Claim 84 was amended to specifically recite that he bandpass filter passes more than the desired

channel and does not form an image rejection filter as does the prior art. Accordingly, as amended, claim 84 should be held allowable.

Claim 98 has been amended to recite that the bandpass filter is constructed in the same substrate as the first mixer. Ash does not teach such a structure and claim 98, as amended, should be held allowable.

Claims 103 and 109 have been amended to recite that the bandpass filter passes at least four channels. Ash does not show such a method of operation. Accordingly, claims 103 and 109, as amended, should beheld allowable.

The Examiner has rejected claims 12 and 17 in view of Ash with the statement that Ash "shows the claimed second mixer which generates the second IF signal." Claims 12 and 17 have been amended to specifically recite that the second IF signal still contains all of the channels of information as contained in the original RF signal. As discussed above, Ash does not teach such a system and accordingly claims 12 and 17 should be held allowable, as amended.

With respect to claim 40, the Examiner has indicated that Ash discloses the step of filtering the second IF signal with a band pass filter. The filter that Ash is discussing is a low pass filter since all of the channels except for the desired channel have already been removed from the signal. Accordingly, in Ash only the remaining noise signals are removed, whereas claim 40 requires the filtering out of certain valid channels. Again, this is a different structure from the Ash reference. Accordingly, it is respectfully suggested that claim 40 is allowable as originally presented and particularly since it depends upon claim 35, which has been discussed above.

The Examiner has rejected claims 8, 41, 42, 44, 45, 48, 50-76, 90 and 114-116 under 35 U.S.C. § 103 as being unpatentable over Ash. With respect to claim 8, the examiner has said that Ash (Fig. 1 reference 68) allows for the use of higher frequency ranges in the neighborhood of 1200 MHz. The Examiner goes on to say

> "It would have been obvious to one skilled in the art at the time the invention was made to use higher frequencies in order to allow the receiver to function within a broader range and allow the user to tune into a wider variety of broadcast signals."

The Examiner's statement is only accurate with respect to a portion of the signals. In Ash in FIGURE 1, element 68 is a UHF VCO circuit which deals with the UHF portion of the signal. There is no IF signal as claimed. Claim 8 on the

other hand is directed to the use of the 1200 MHz signal in the IF frequency after the first mixer, while in Ash the 1200 MHz signal is used in the first mixer. The Ash circuit is a different structure and accordingly, claim 8 should be allowable in view of the Ash reference.

The Examiner goes on to say as for claims 41, 42 and 51-54, the use of various filtering techniques at different frequencies is commonly known in the art of signal processing and would have been obvious to one skilled in the art at the time the invention was made in order to achieve the desired result.

The problem with the Examiner's statement is that the desired result, i.e., the ability to do the entire signal detection on a single substrate by doing portions of the filtering at different stages is not taught in any of the prior art. In addition, the precise structure claimed in the above-identified claims is not taught by Ash. While it is true that various filtering techniques at different frequencies are commonly known in the art, the invention as recited in the claims, is directed to a specific structure and method of operation where the frequency spectrum is filtered at different places in the process to achieve a different goal from past circuits, i.e., detection on a single substrate. However, while the same overall end result is achieved, it is the method of achieving the result which has heretofore not been known. Accordingly, the Examiner's rejection under 35 U.S.C. 102 and 103 with respect to the Ash patent is respectfully traversed.

With respect to claim 58, Applicant does not claim that the filtering and mixing is new, only that in the combination recited with respect to the other elements taken as a whole, the prior art does not disclose such a system nor is such a system rendered obvious by the art cited by the Examiner.

The Examiner has rejected claims 3, 14-16, 19-23, 28, 34, 38 and 49 under 35 U.S.C. as being unpatentable over Ash in view of Mau. With respect thereto, the Examiner has stated "Mau (Col. 4, lines 44-47) discloses a second mixer which is an image rejection mixer."

In this regard the Examiner's comments are not well taken. Mau does not disclose an image rejection mixer but discloses an image tracking filter 28 shown in Fig. 2 of Mau. The filter of Mau functions in a totally different manner than the claimed second mixer of the instant invention. In fact, in Mau the image tracking filter is specifically identified as being in front of the mixer whereas in the claimed invention the image rejection mixer is the second mixer. These are totally different

concepts operating in a totally different manner. Accordingly, it is respectfully submitted that the Examiner's rejection of the claims based upon the combination of Ash and Mau is not well founded.

With respect to claim 22, the Examiner has stated:

> "Ash illustrates a television receiver wherein the second operating frequency is less than the first predetermined frequency and the frequency of the image signal rejected by the second mixer is less than the second operating frequency by an amount equal to the second predetermined frequency (Ash, Fig. 2)".

Since Ash rejects all channels, other than the desired channel, prior to the second mixer, there is no comparison between the Ash reference and the claim. Accordingly, with respect to the rejection of the claims based upon 35 U.S.C. § 103 in view of Ash and Mau, the Examiner's rejection is respectfully traversed.

Claims 77-81 are rejected under 35 U.S.C. as being unpatentable over Ash in view of Nordholt. Claims 82 and 83 were rejected under 35 U.S.C. § 103 as being unpatentable over Ash in view of Nordholt and further in view of Sussman. These rejections will not be discussed since it is believed that the claims are dependent from claims that would otherwise be allowable, as discussed above, and accordingly are deemed to be allowable as presented. Neither the Nordholt nor the Sussman reference teaches a system or method for decoding a TV signal in the manner claimed. Each of these references are conventional decoders and illustrate the problem with the prior art.

Claim 117 has been added directed to the manner in which the desired channels are removed from the RF signal in a manner which would make the circuit implementable on a single substrate. None of the art cited renders claim 117 anticipated or obvious for the reasons discussed above with respect to claim 1. The manner and order for performing the steps of claim 117 which are not shown in the prior art, allow the circuit to be constructed, if desired, on a single substrate. Accordingly, it is respectfully submitted that newly added claim 117 is allowable as well as claims 118 and 119 which depend therefrom.

For the above reasons, the Applicant submits that the claimed design is of patentable merit under both 35 U.S.C. § 102 and 35 U.S.C. § 103. Accordingly, the Applicant submits that this application is in full condition for allowance and asks that the Examiner pass this case to issue.

Applicant's attorney respectfully requests that the Examiner call him at the below listed number if the Examiner believes that such a discussion would be helpful in resolving any remaining problems.

Respectfully submitted,

FULBRIGHT & JAWORSKI L.L.P.
Attorneys for Applicant

Date: May 8, 1996

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(214) 855-8333

05/07/96
0390245

ATTORNEY DOCKET No. 47681-P… …US-954919

PATENT APPLICATION

IN THE U.S. PATENT AND TRADEMARK OFFICE

MAIL ROOM MAY 13 1996 PAT. & TRADEMARK OFFICE

Inventor: Robert Rudolf Rotzoll

Serial No.: 08/426,680

Filing Date: April 21, 1995

Examiner: Johnson

Group Art Unit: 2602

Title: MONOLITHIC TELEVISION TUNER

Assistant Commissioner
for Patents
Washington, D.C. 20231

Date of Deposit May 8, 1996

I hereby certify that this is being deposited with the United States Postal Service as First Class Mail in an envelope addressed to: Assistant Commissioner for Patents, Washington, D.C. 20231.

By: Elva Wilson

Typed Name: Elva Wilson

Transmittal of Amendment

Sir:

Transmitted herewith is/are the following in the above-identified application:

(x) Response/Amendment

(x) New fee as calculated below

( ) No additional fee (Address envelope to "Box NON-FEE Amendment")

( ) Other: ______

CLAIMS AS AMENDED BY OTHER THAN A SMALL ENTITY						
(1) FOR	(2) CLAIMS REMAINING AFTER AMENDMENT	(3) NUMBER EXTRA	(4) HIGHEST NUMBER PREVIOUSLY PAID FOR	(5) PRESENT EXTRA	(6) RATE	(7) ADDITIONAL FEES
TOTAL CLAIMS	119	MINUS	116	= 3	X $22	$ 66.00
INDEP. CLAIMS	10	MINUS	9	= 1	X $78	$ 78.00
[ ] FIRST PRESENTATION OF A MULTIPLE DEPENDENT CLAIM					+$ 250	$
					TOTAL ADDITIONAL FEE FOR THIS AMENDMENT	$ 144.00

Enclosed is a check in the amount of $144.00. At any time during the pendency of this application, please charge any fees required or credit any overpayment to Deposit Account 06-2380 pursuant to 37 CFR 1.25. Additionally, please charge any fees to Deposit Account 06-2380 under 37 CFR 1.19, 1.20 and 1.21. A duplicate copy of this sheet is enclosed.

Respectfully submitted,

FULBRIGHT & JAWORSKI L.L.P.
Attorneys for Applicant

Date: May 8, 1996

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Notes:

TUNE 048285

TAB 5

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PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application:	Robert Rudolf Rotzoll
Serial No.:	08/426,020
Filed:	April 21, 1995
Art Unit:	2602
Examiner:	Sherrie Hsia
For:	MONOLITHIC TELEVISION TUNER

Assistant Commissioner
for Patents
Washington, D.C. 20231

I hereby certify that this correspondence is, on the date shown below, being deposited with the United States Postal Service with sufficient postage as first class mail, in an envelope addressed to the Assistant Commissioner for Patents, Washington, DC 20231.

Date deposited: October 1, 1996

Signature: Linda Cuccia

Printed Name: LINDA CUCCIA

AMENDMENT

Dear Sir:

In response to the Office Action, mailed August 27, 1996, Applicant respectfully requests that the Examiner reconsider the application in view of the amendments and comments set forth below.

In the Title

Delete the Title of the invention and replace with -- HIGHLY INTEGRATED TELEVISION TUNER ON A SINGLE MICROCIRCUIT --.

In the Specification

On page 11, please delete the descriptions of Figures 1 and 2 and insert the following descriptions therefor:

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-- FIGURE 1 is a high level block diagram of the functional components of the present invention;

FIGURE 2 is a pin layout diagram of an integrated circuit in accordance with the present invention; --

B1

On page 18, line 33, delete "100" and insert -- 200 -- after "circuit".

On page 24, line 16, delete "420" and insert -- 422 -- after "IFAMP"; and in line 28, delete "audio delay line (ADLY)" and insert -- audio phase shifter (APS) -- after "and".

On page 28, lines 20-29 should be amended to read as follows:

B2

trimming techniques to calibrate MIX2. As shown in FIGURE 6, t[T]he 50dB image rejection may be achieved by paralleling two Gilbert cell multipliers 610 and 612 at MIX2 input and using fairly high-order balanced RC-phase shifters [611,] 613, 614 with very low

On page 38, line 8, delete "470" and insert -- 170 -- after "interface".

In the Claims

Please amend the claims as shown.

In Claim 12, line 4, delete "the original" and insert -- said -- after "in".

B3

~~17.~~ 16. (Twice Amended) The television receiver as set forth in Claim 1 further comprising a first IF signal generated by said first mixer wherein said second mixer mixes said second reference signal and said first IF signal to generate a second IF signal containing all of the channels of information as contained in [the original] an RF signal received from said RF signal source.

In Claim 21, line 2, change "the" to -- a --.

Sub C4

B4

28 ~~29~~. (Twice Amended) A method of processing a received RF signal, the method comprising the steps of:

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said [IF] mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal.

In Claim 34, line 5, change "second" to -- first --.

B5

34 ~~35~~. (Twice Amended) A method of processing a received RF signal, the method comprising the steps of:

filtering all frequency components above an input cutoff frequency out of the received RF signal to thereby produce a first filtered signal;

amplifying the first filtered signal;

subtractively mixing the amplified first filtered signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said [IF] mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal.

In Claim 38, line 5, change "second" to -- first --.

B6

38 ~~39~~. (Amended) The method as set forth in Claim 34 ~~35~~ wherein the step of mixing the first IF signal with the second reference signal is a down conversion mixing, wherein [the] a selected carrier signal appearing in the first IF signal at [the] a first predetermined frequency appears in the second IF signal at a second predetermined frequency and wherein the second predetermined frequency is lower than the first IF signal.

In Claim 42, line 1, change "35" to -- 40 --.

In Claim 47, line 3, insert -- and amplified -- after "filtered".

In Claim 49, line 3, change "received" to -- outputted --.

In Claim 52, line 2, change "51" to -- 43 --.

In Claim 53, line 3, change "the" to -- a --.

In Claim 54, line 2, change "video" to -- visual --.

57 ~~58~~. (Amended) The television receiver as set forth in Claim 55 ~~56~~ wherein said television receiver further comprises:

B7 a [forth] third filter coupled to an output of said second amplifier, wherein said [fourth] third filter is a bandpass filter for extracting a video carrier signal from an output signal of said second amplifier; and

a third mixer having a first input and a second input, wherein said first input of said third mixer is coupled to said output of said second amplifier and said second input of said third mixer is coupled to an output of said [fourth] third filter and wherein an output signal on an output of said third mixer is a video baseband signal.

In Claim 59, line 3, change "fourth" to -- third --.

In Claim 60, line 6, change "fourth" to -- third --.

In Claim 62, line 3, change "fifth" to -- fourth --.

74 ~~75~~. (Amended) The television receiver as set forth in Claim [52] 55 ~~56~~ wherein

B8C said television receiver further comprises a [sixth] third ~~fifth~~ filter coupled to an output of said second amplifier, wherein said [sixth] third ~~fifth~~ filter is a narrow-band filter for

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B8 CONT

filtering an audio component signal from a video output signal on said output of said second amplifier.

In Claim 76, line 2, change "sixth" to -- fifth --.

In Claim 77, line 4, change "sixth" to -- fifth --; and in line 7, change "sixth" to -- fifth --.

In Claim 78, line 6, change "sixth" to -- fifth --.

In Claim 79, line 3, change "seventh" to -- sixth --; and in line 4, change "seventh" to -- sixth --.

In Claim 86, line 4, change "received" to -- outputted --.

In Claim 88, line 3, change "received" to -- outputted --.

In Claim 99, line 3, change "received" to -- outputted --.

In Claim 107, line 7, delete "the RF signal" and insert -- the selected carrier frequency -- after "than".

B9c

~~111~~ 110. (Amended) The method as set forth in Claim ~~110~~ 109 wherein the step of mixing the [RF] amplified first filtered signal with the first reference signal is an up-conversion mixing, wherein a selected carrier signal in ~~the~~ said received RF signal appears in the first IF signal at a first predetermined frequency and wherein said first predetermined frequency is higher than the [RF signal] selected carrier frequency.

In Claim 113, line 4, delete "the" each time it occurs and insert -- a -- after "wherein" and insert -- a -- after "at".

116. ~~117.~~ (Amended) The method of extracting a single communications channel from a large number of such channels wherein each such channel has a defined frequency range, and wherein the large number of channels are communicated within an upper and a lower frequency boundary, the method comprising the steps of:

from an input signal containing all of the large number of channels removing all frequencies greater than the upper frequency boundary;

translating all of the channels contained within the input signal to a translated frequency range such that a desired single channel to be extracted from the input signal is positioned centrally within the translated frequency range, the translated frequency range being positioned at frequencies higher [in the frequency spectrum] than the upper frequency boundary;

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removing from the translated input signal all channels extending from the center of the translated frequency range four channels or more in each frequency direction;

translating the remaining channels within the input signal to a second translated frequency range, such that the desired single channel to be extracted from the input signal is positioned centrally within the second translated frequency range having a low end frequency boundary defined as high enough not to cause phase distortion in any subsequent filtering and a high end frequency boundary defined as a [the] frequency beyond which a filter with proper bandwidth, noise and distortion can be integrated in a substrate; and

removal of all remaining frequencies which are not part of the desired single channel.

In Claim 118, line 3, insert -- single -- after "desired".

Remarks

Claims 1-119 are pending in the application. Claims 1-11, 24-27, 43-51, 56-74, 84-112 and 114-116 have been allowed by the Examiner. Claims 12-23, 28-42, 52-55, 75-83, 113, and 117-119 have been rejected by the Examiner.

The issues outstanding in the instant application are as follows:

-- The Title has been rejected as not descriptive under MPEP § 606.01.

-- The disclosure is objected to because of various informalities.

-- The Drawings are objected to and a proposed drawing correction is required.

-- Claims 12-23, 28-42, 52-55, 75-83, 113 and 117-119 are rejected under 35 U.S.C. § 112, second paragraph, as being indefinite for failing to particularly point out and distinctly claim the subject matter which applicant regards as the invention.

-- Claims 47, 49, 58, 60, 62, 63, 75-80, 86, 88, 99, 107, 111 and 118 are objected to because of various informalities.

Applicant hereby traverses all the outstanding rejections and objections, and requests reconsideration and withdrawal in light of the amendments and remarks contained herein.

Objection to the Title

The Examiner has rejected the title of the invention as not descriptive of the invention as required under MPEP § 606.01. Applicant has amended the title of the invention to be more descriptive. Applicant submits that the amended title satisfies the requirements of MPEP § 606.01 and is allowable.

Objection to the Specification

The Examiner has noted the length of the specification and the resulting difficulty in detecting all minor errors. Applicant has reviewed the specification for errors and has amended the Brief Description of the Drawings to correctly identify the drawings in FIGURES 1 and 2. Additionally, Applicant has amended the Detailed Description of the Invention at page 28 to identify element 611 as a phase splitter.

The Examiner has objected to the disclosure as failing to refer to FIGURE 6 in the Detailed Description of the Invention, and because various elements in the disclosure are mislabeled. Applicant has amended the disclosure to add a specific reference to FIGURE 6. Additionally, Applicant has amended the disclosure to correctly identify the elements cited by the Examiner.

Objection to the Drawings

The Examiner has objected to the drawings. Specifically, in FIGURE 6, the Examiner states that "splitter" should be "shifter". The Examiner has also required a proposed drawing correction in response to this objection. Applicant submits that FIGURE 6 is not in error in view of the amendments to the specification. Applicant has amended the specification to identify element 611 as a phase splitter. As disclosed in the specification at page 22, lines 26-29, element 611 is a phase splitter which splits the local oscillator signal from LO2 into two phase quadrature signals. Applicant submits that the drawings are not in error and are in condition for allowance.

Rejection under 35 U.S.C. § 112, second paragraph

Claims 12-23, 28-42, 52-55, 75-83, 113 and 117-119 have been rejected under 35 U.S.C. § 112, second paragraph, as indefinite for failing to particularly point out and distinctly claim the subject matter which applicant regards as the invention.

Applicant has amended Claims 12, 17, 21, 29, 35, 39, 42, 53, 54, 75, 113 and 117 to provide antecedent basis for the various elements identified by the Examiner. Applicant has amended Claim 52 to depend from Claim 43 instead of Claim 51. Additionally, as requested by the Examiner, Applicant has reviewed all the terms in the claims for proper antecedent basis and proper word usage. Applicant submits that Claims 12, 17, 21, 29, 35, 39, 42, 53, 54, 75, 113 and 117, as amended, particularly point out and distinctly claim the subject matter of the invention and are allowable under 35 U.S.C. § 112, second paragraph.

Applicant traverses the Examiner's rejection of Claims 14, 19, 28, 34 and 38 as confusing. The Examiner states that it is not understandable why the second mixer rejects the image signal of the second reference signal or of the first IF signal

from the second IF signal since the second mixer generates the second IF signal. The Examiner further states that it appears that the second IF signal should be the first IF signal or the second reference signal. This characterization of the second IF signal is not correct. Applicant discloses the basic mixer formula on page 20 of the specification. At line 18, Applicant discloses the equation for determining the frequency of the mixer output signal:

$$f_{IF} = |f_{RF} \pm f_{LO}|$$

It can be seen from the formula that the frequency of the mixer output signal (f_{IF}) is dependent upon the frequencies of the mixer input signals (f_{RF} and f_{LO}). In the case of MIX2, as illustrated on page 22, lines 9-22, f_{RF} is the frequency of the first IF signal and f_{LO} is the frequency of the second reference signal. In the preferred embodiment, the desired frequency of the first IF signal is 1200 Mhz and the desired frequency of the second reference signal is 1180 MHz. Applying the mixer equation above it can be seen that the output from MIX2 contains signals at frequencies of 20 MHz (|1200 - 1180|) and 2380 MHz (|1200 + 1180|). The specification discloses that in the preferred embodiment the 20 MHz signal is used for further processing.

In actual implementation of the invention, it is possible that undesired signals at interfering frequencies will also be present in the first IF signal or in the second reference signal. The purpose of the image rejection mixer is to prevent these extraneous frequencies from interfering with the desired 20 MHz signal. The mixer rejects input signals at frequencies which could produce an undesired 20 MHz output. For example if the second reference signal contains an extraneous signal at 1220 MHz, in addition to the desired signal at 1180 MHz, mixing of the 1220 MHz signal in MIX2 will produce an undesired output signal at 20 MHz. (|1200 - 1220| = 20 MHz.) By rejecting this 1220 MHz image signal associated with the second reference signal, MIX2 will prevent the undesired output signal from propagating further through the circuit.

Extraneous frequencies in the first IF signal can also produce an undesired signal at 20 MHz in the output of MIX2. If the first IF signal contains a signal at 1160 MHz, the output of MIX2 would contain an undesired signal at 20 MHz. (|1160 - 1180| = 20 MHz.) MIX2 must reject these image signals associated with the first IF signal.

Claims 14, 19, 28, 34 and 38 are directed to eliminating undesired signals in the output of MIX2 as discussed above. Claims 34 and 38 have been amended to clarify that the image signals rejected in those claims are associated with the first IF signal. In view of the operation of MIX2 and the undesired effects of the image signals on the MIX2 output as discussed above, Claims 14, 19, 28, 34 and 38 are not confusing. Applicant submits that Claims 14, 19, 28, 34 and 38, as amended, particularly point out and distinctly claim the subject matter of the invention and are allowable under 35 U.S.C. § 112, second paragraph.

Objection due to informalities

The Examiner has objected to Claims 47, 49, 58-60, 62, 63, 75-80, 86, 88, 99, 107, 111 and 118 because of various informalities.

Applicant has amended Claims 47, 49, 86, 88, 99, 107, 111 and 118 in accordance with the Examiner's recommendations to overcome the specified informalities. Additionally, Applicant has amended Claims 58-60, 62 and 75-79 by renaming certain filters. Applicant has labeled the filters so that they are consistently identified across all the claims.

Applicant believes that the Examiner has improperly identified Claim 63 as requiring correction. Line 4 of Claim 63 does not contain a reference to a fifth filter as stated by the Examiner. Therefore, Claim 63 has not been amended.

Conclusion

The Examiner has stated that Claims 29-42 and 117-119 would allowable if rewritten or amended to overcome the rejections under 35 U.S.C. § 112. The Examiner has further stated that Claims 12-23, 28, 52-55, 75-83 and 113 would be allowable if rewritten to overcome the rejections under 35 U.S.C. 112 and to include all of the limitations of the base claim and any intervening claims.

Applicant has amended Claims 12, 17, 21, 29, 34, 35, 38, 39, 42, 47, 49, 52, 53, 54, 58-60, 62, 75-79, 86, 88, 99, 107, 111, 113, 117 and 118. Applicant submits that the amendments overcome the Examiner's rejections under 35 U.S.C. § 112 and that the amended claims and all claims depending therefrom are allowable.

For the above reasons, the Applicant submits that the claimed design is of patentable merit under 35 U.S.C. § 112, second paragraph. Accordingly, the Applicant submits that this application is in full condition for allowance without introducing any new issues.

Applicant's attorney respectfully requests that the Examiner call him at the below listed number if the Examiner believes that such a discussion would be helpful in resolving any remaining problems.

Respectfully submitted,

FULBRIGHT & JAWORSKI L.L.P.
Attorneys for Applicant

Date: 10-1-96

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09/27/96

ATTORNEY DOCKET NO. 4.,81-P023US-954919 **PATENT APPLICATION**

IN THE U.S. PATENT AND TRADEMARK OFFICE

Inventor(s):	Robert Rudolf Rotzoll		
Serial No.:	08/426-080	Examiner:	Sherrie Hsia
Filing Date:	April 21, 1995	Group Art Unit:	2602
Title:	MONOLITHIC TELEVISION TUNER		

ASSISTANT COMMISSIONER
FOR PATENTS
Washington, D.C. 20231

TRANSMITTAL LETTER FOR AMENDMENT

Sir:

Transmitted herewith is/are the following in the above-identified application:

(X)	Response/Amendment	( )	Petition to extend time to respond
( )	New fee as calculated below	( )	Supplemental Declaration
(X)	No additional fee (Address envelope to "Box NON-FEE Amendment")		
( )	Other: ______		

CLAIMS AS AMENDED BY OTHER THAN A SMALL ENTITY						
(1) FOR	(2) CLAIMS REMAINING AFTER AMENDMENT	(3) NUMBER EXTRA	(4) HIGHEST NUMBER PREVIOUSLY PAID FOR	(5) PRESENT EXTRA	(6) RATE	(7) ADDITIONAL FEES
TOTAL CLAIMS	37	MINUS	37	= 0	X $22	$ 00.00
INDEP. CLAIMS	4	MINUS	4	= 0	X $80	$ 00.00
[ ] FIRST PRESENTATION OF A MULTIPLE DEPENDENT CLAIM					+$250	$
EXTENSION FEE	1ST MONTH $110	2ND MONTH $390	3RD MONTH $930	4TH MONTH $1470		$
					TOTAL ADDITIONAL FEE FOR THIS AMENDMENT	$ 00.00

Enclosed is a check in the amount of $00.00. At any time during the pendency of this application, please charge any fees required or credit any overpayment to Deposit Account 06-2380 pursuant to 37 CFR 1.25. Additionally, please charge any fees to Deposit Account 06-2380 under 37 CFR 1.19, 1.20 and 1.21. A duplicate copy of this sheet is enclosed.

I hereby certify that this correspondence is being deposited with the United States Postal Service as first class mail in an envelope addressed to: Assistant Commissioner for Patents, Washington, D.C. 20231

Date of Deposit: 10-1-96

Typed Name: LINDA CUCCIA

Signature: Linda Cuccia

Respectfully submitted,

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Date: 10-1-96

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Notes:

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TAB 6

#9 Amdt/C

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PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application of:	ROBERT RUDOLF ROTZOLL
Serial No.:	08/426,080
Filed:	April 21, 1995
Art Unit:	2602
Examiner:	Sherrie Hsia
For:	MONOLITHIC TELEVISION TUNER

Assistant Commissioner
for Patents
Washington, D.C. 20231

I hereby certify that this correspondence is, on the date shown below, being deposited with the United States Postal Service with sufficient postage as first class mail, in an envelope addressed to the Assistant Commissioner for Patents, Washington, DC 20231.

Date deposited: March 12, 1997

Signature: Elva Wilson

Printed Name: Elva Wilson

AMENDMENT

Dear Sir:

In response to the Office Action, mailed January 09, 1997, Applicant respectfully requests that the Examiner reconsider the application in view of the amendments and comments set forth below.

In the Specification

Please amend the Specification as follows:

On page 28, lines 25-26 should be amended to read as follows:

C1 "... less than 0.2 degrees over a 20 MHz bandwidth. Multipliers 610 and 612 receive inputs from Phase Splitter 611 and MIX1 408. These are traditionally avoided due to relatively high noise ..."

In the Claims

Please amend the Claims as follows:

C2

1. (Twice Amended) A television receiver comprising:

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled directly to an output of said first mixer and a second input coupled to said second reference signal, wherein said second mixer is an image rejection mixer.

Delete Claim 3.

23 ~~24~~. (Twice Amended) A television receiver comprising:

a receiver input coupled to an RF signal source;

D3

a first mixer physically located on an integrated circuit substrate having a first input coupled to said receiver input and a second input coupled to a first reference signal having a first operating frequency, wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value; and

a second mixer physically located on the same integrated circuit substrate as the first mixer and having a first input directly coupled without leaving the substrate to an output of said first mixer and a second input coupled to a second reference signal having a second operating frequency, wherein said second operating frequency is fixed.

28 ~~29~~. (Third Amendment) A method of processing a received RF signal, the method comprising the steps of:

24

mixing the RF signal with a first reference signal having a first operating frequency to thereby produce a first IF signal; and

mixing the first IF signal with a second reference signal having a second operating frequency to thereby produce a second IF signal, said mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal, wherein said RF signal mixing and said first IF signal mixing operations are accomplished on the same integrated circuit substrate.

42 ~~43~~. (Twice Amended) A television receiver comprising:

25

an input filter coupled to an RF signal source;

a first amplifier coupled to an output of said input filter;

a first mixer having a first input and a second input, wherein said first input of said first mixer is coupled to an output of said first amplifier;

a first oscillator coupled to said second input of said first mixer, wherein said first oscillator generates a first reference signal having a first operating frequency;

a second mixer having a first input and a second input, wherein said first input of said second mixer is coupled directly to an output of said first mixer without the connection of a [saw] filter therebetween;

a second oscillator coupled to said second input of said second mixer, wherein said second oscillator generates a second reference signal having a second operating frequency; and

a second filter coupled to an output of said second mixer.

In Claim 62, line 4, change "fifth" to --fourth--.

In Claim 63, line 4, after "signal" add --from said fourth filter--.

In Claim 75, lines 2 and 3, change "fifth" to --third--.

In Claim 76, line 2, change "fifth" to --third--.

76 ~~77.~~ (Twice Amended) The television receiver as set forth in Claim 74 ~~75~~ wherein said television receiver further comprises:

C6

audio phase detector circuitry coupled to an output of said [fifth] third filter; and

a [fourth] third mixer having a first input and a second input, wherein said first input of said [fourth] third mixer is coupled to said output of said [fifth] third filter and said second input of said [fourth] third mixer is coupled to an output of said audio phase detector circuitry.

In claim 78, line 6, change "fifth" to --third--.

78 ~~79.~~ (Twice Amended) The television receiver as set forth in Claim 76 ~~77~~ wherein said television receiver further comprises a [sixth] fourth filter coupled to an output of said [fourth] third mixer, wherein said [sixth] fourth filter is a low-pass filter with a cut-off frequency of 120 Kilohertz.

C7

In claim 80, line 4, change "fourth" to --third--.

83 ~~84~~ (Twice Amended) A television receiver comprising:

C8

a receiver input coupled to an RF signal source;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to said receiver input and a second input coupled to said first reference signal;

a first bandpass filter coupled to an output of said first mixer; the bandpass filter passing more than one channel and [not forming an] performing partial image rejection as well as limiting overall signal power levels to be processed by [for a] subsequent [mixer] circuitry;

a second reference signal having a second operating frequency; and

[a second] an image rejection mixer having a first input coupled to an output of said first bandpass filter and a second input coupled to said second reference signal.

In claim 111, line 3, delete "the" before "RF signal" and replace with --said received--.

Remarks

Claims 1-119 are pending in the application. Claims 1-57 and 84-119 have been rejected and claims 58-83 are objected to in the Office Action. Therefore, claims 1-119 are at issue in the application. The outstanding issues in the instant application are:

-- Claims 1-11, 24-27, 43-51, 56-74, 84-112 and 114-116, which were noted as allowable in the August 27, 1996 Office Action, now stand rejected.

-- The Drawings are objected to for failure to mention element 611 in the Specification.

-- Claims 62, 63, 75-80 and 111 are objected to for informalities.

-- Claim 42 is rejected under 35 U.S.C. § 112, second paragraph.

-- Claims 1, 2, 4-7, 9, 12, 13, 17, 18, 24-27 and 29 are rejected under 35 U.S.C. § 102(b) as being anticipated by U.S. Patent No. 4,855,835 to Tobita ("Tobita").

-- Claims 24-27, 29-33, 35-37, 39, 40, 43, 46, 47, 56, 84-89, 91-114 and 117-119 are rejected under 35 U.S.C. § 102(b) as being anticipated by U.S. Patent No. 4,340,975 to Onishi et al. ("Onishi").

-- Claims 3, 8, 10, 11, 14-16, 19-23, 28 and 30-34 are rejected under 35 U.S.C. § 103(a) as being unpatentable over Tobita in view of Onishi.

-- Claims 38, 41, 42, 44, 45, 48-55, 57, 90, 115 and 116 are rejected under 35 U.S.C. § 103(a) as being unpatentable over Onishi.

-- Claims 58-83 are objected to as being dependent from a rejected base claim.

Applicant hereby traverses all outstanding rejections and objections and requests reconsideration and withdrawal in view of the amendments and remarks contained herein.

Claims previously noted as allowable

In the previous Office Action, dated August 27, 1996, the Examiner indicated that claims 1-11, 24-27, 43-51, 56-74, 84-112 and 114-116 were allowed. The Examiner has now rejected claims 1-11, 24-27, 43-51, 56, 57, 84-112 and 114-116, among others. Applicant notes that Form Paragraph 7.50 and MPEP § 706.04 are

not cited in the current Office Action and that there is no indication that a primary examiner approved the rejection of the previously allowed claims.

Objection to the Drawings

Applicant has amended the Specification to include a reference to element 611 of Figure 6. No new matter is added by this amendment. Applicant submits that this amendment brings the Drawings and Specification into compliance with 37 CFR 1.84(p)(5) and that no further corrections to the Drawings are required.

Objections for informalities

Applicant has amended claims 62, 63, 75-80 in accordance with the Examiner's suggestions to clarify the references to the mixers and filters in those claims. Applicant has amended claim 111 to clarify that the selected carrier signal, which appears in the first IF signal at a first predetermined frequency, is related to the received RF signal, not the amplified first filtered signal as proposed in the Office Action. Applicant submits that the above amendments overcome the objections cited in the Office Action and that the claims are in condition for allowance.

Rejection under 35 U.S.C. § 112, second paragraph

Claim 42 is rejected as being indefinite for failing to recite sufficient antecedent basis for "the bandpass filter" limitation. Claim 42, as amended by Applicant's October 1, 1996 Amendment, depends from claim 40 and thereby includes all the limitations of claim 40. Claim 40 recites the limitation "a bandpass filter," thereby providing antecedent basis for dependent claim 42. Applicant submits that claim 42 complies with 35 U.S.C. § 112 and is in condition for allowance.

Rejections under 35 U.S.C. § 102(b)

Tobita Reference

Claims 1, 2, 4-7, 9, 12, 13, 17, 18, 24-27 and 29 stand rejected under 35 U.S.C. § 102(b) as being anticipated by Tobita.

Claim 1, as amended, includes the limitation that the second mixer is an image rejection mixer. MPEP § 2131 requires that a reference must teach every

element of a claim to anticipate that claim. As admitted in the Office Action, Tobita does not disclose an image rejection mixer, therefore, the Tobita reference cannot anticipate claim 1 under 35 U.S.C. § 102.

Claim 24 includes the limitations that the first mixer is located on an integrated circuit substrate, and that the second mixer is located on the same integrated circuit substrate. Claim 29, as amended, also requires that the mixing operations take place on the same integrated circuit substrate. Contrary to the statements in the Office Action, there is no teaching in Tobita that the first and second mixers are located on the same integrated circuit substrate. Furthermore, Applicant does not find any mention of an integrated circuit substrate in the Tobita disclosure. Therefore, Tobita does not disclose every element of claims 24 or 29 as required by MPEP § 2131.

Independent claims 1, 24 and 29 contain limitations which are not disclosed or even hinted at in the Tobita reference. Therefore, Tobita can not anticipate, nor render obvious, these claims. Applicant submits that claims 1, 24 and 29 are allowable under 35 U.S.C. § 102(b) over the Tobita reference.

Claims 2, 4-7, 9, 12, 13, 17, 18 and 25-27 depend from claims 1, 24 or 29 and thereby include all of the limitations of the independent base claim. Since Tobita does not anticipate claims 1, 24 or 29, it cannot be used to anticipate the dependent claims. Applicant submits that claims 2, 4-7, 9, 12, 13, 17, 18 and 25-27 are not anticipated by Tobita and are in condition for allowance.

Onishi Reference

Claims 24-27, 29-33, 35-37, 39, 40, 43, 46, 47, 56, 84-89, 91-114 and 117-119 stand rejected under 35 U.S.C. § 102(b) as being anticipated by Onishi.

Claim 24, as amended, requires that the first input of the second mixer be "directly coupled without leaving the substrate to an output of [the] first mixer." Onishi does not teach that the input of the second mixer can be directly coupled to the first mixer. Instead, Onishi requires that a bandpass filter be placed between the two mixers. MPEP § 2131 requires that the elements of an anticipating reference must be arranged as required by the claim. The circuit of Onishi cannot anticipate claim 24 because it is not arranged to include the limitation of direct coupling between the mixers. Onishi also fails to teach that the "second mixer is

physically located on the same integrated circuit substrate as the first mixer" as required by claim 24. Onishi discloses circuit elements formed on separate and discrete integrated circuits (description of Figures 9-18). Since the main portions of the circuit are on separate integrated circuits, each having "terminals" to connect the elements together, Onishi also fails to disclose or even suggest a circuit in which the first input of the second mixer is coupled to the output of the first mixer "without leaving the substrate." Onishi does not teach every element of claim 24, therefore, it cannot anticipate, nor render obvious, the claim.

Claims 29 and 35 require "mixing of the first IF signal with the second reference signal occurring prior to eliminating any channels from the first IF signal." Each embodiment of Onishi clearly requires the first IF signal to pass through a band pass filter (27, 48 or 48′) prior to mixing with the second reference signal. The band pass filter is used to eliminate channels adjacent to the first intermediate frequency (column 10, lines 51-59). Onishi not only fails to teach all of the required elements of claims 29 and 35, but it also explicitly teaches away from the "prior to eliminating any channels" limitation of claims 29 and 35.

Claim 43, as amended, requires that an "input of [the] second mixer is directly coupled to an output of [the] first mixer without the connection of a filter therebetween." As stated above, each embodiment of Onishi clearly requires the output of the first mixer to connect to a band pass filter (27, 48 or 48′) before connecting to the second mixer input. Therefore, Onishi does not teach each element of claim 43.

Claim 84 requires that the bandpass filter pass more than one channel. On the other hand, the bandpass filters of Onishi has a bandwidth of 5 MHz and they are designed "so that interferences by adjacent channels in the following circuits are prevented to the utmost." (Column 10, lines 57-59). Furthermore, claim 84 requires that the output of the bandpass filter be coupled to an image rejection mixer. Onishi does not teach that the second mixer is an image rejection mixer. In fact, Onishi teaches against using an image rejection mixer attached to the output of the bandpass filter, because "only the fixed frequency bandpass filter is used to reject the image frequency." (Column 12, lines 2-3). Therefore, Onishi does not teach the elements of claim 84.

Claim 98 requires "a first bandpass filter constructed in the same substrate as [the] first mixer." Onishi discloses circuit elements formed on separate and discrete integrated circuits (description of Figures 9-18). Furthermore, the band pass filter 48 is not constructed of a single substrate itself. Instead filter 48 of Onishi is constructed of at least two dielectric substrates which are separated by shielding plates (column 19, line 61 - column 20, line 9). Thus, Onishi does not teach each and every element of claim 98.

Claims 103 and 109 require "filtering the first IF signal in a first bandpass filter to produce an output passband signal having at least four channels." Claim 117 requires removing all channels which extend four channels or more from a desired single channel in a translated signal. The remaining channels are then translated to a second translated frequency range. Therefore, claims 103, 109 and 117 require mixing or translating multiple channels (and not a single channel) in the second mixing or translating operation. Onishi teaches that the band pass filter removes all channels except the desired channel from the first IF signal. Therefore, Onishi teaches that only one channel can be mixed in the second mixer. Onishi does not teach each and every element of claims 103, 109 or 117.

Onishi does not teach each element of claims 24, 29, 35, 43, 84, 98, 103, 109 or 117. Furthermore, the disclosure of Onishi teaches away from the limitations in the claims. Therefore, Onishi cannot anticipate claims 24, 29, 35, 43, 84, 98, 103, 109 or 117 under 35 U.S.C. § 102(b). Applicant therefore respectfully submits that the claims, as amended, are novel and non-obvious over Onishi and are in condition for allowance.

Claims 25-27, 30-33, 36, 37, 39, 40, 46, 47, 56, 85-89, 91-97, 99-102, 104- 108, 110-114, 118 and 119 depend from the independent claims 24, 29, 35, 43, 84, 98, 103, 109 or 117 and thereby include all of the limitations of the respective base claim. Since Onishi does not anticipate the independent base claims, it cannot anticipate the dependent claims. Applicant submits that claims 25-27, 30-33, 36, 37, 39, 40, 46, 47, 56, 85-89, 91-97, 99-102, 104- 108, 110-114, 118 and 119 are not anticipated, nor rendered obvious, by Tobita and are in condition for allowance under 35 U.S.C. § 102(b).

Rejections under 35 U.S.C. § 103(a)

Combination of Tobita and Onishi

Claims 3, 8, 10, 11, 14-16, 19-23, 28 and 30-34 stand rejected under 35 U.S.C. § 103(a) as being unpatentable over Tobita in view of Onishi.

MPEP § 2142 sets forth three criteria to establish a prima facie case of obviousness: (1) there must be some suggestion or motivation in the references to combine the references, (2) there must be a reasonable expectation of success and (3) the references must teach or suggest all the claim limitations. Applicant submits that the cited references, either alone or combined, do not satisfy these criteria.

The Office Action states that it would have been obvious to one skilled in the art to add Onishi's input filter to the device of Tobita to eliminate the interference signal. Applicant submits that this does not provide sufficient suggestion or motivation for the combination of the automatic fine tuning circuit of Tobita with the VHF-UHF all channel tuner of Onishi. The fact that references can be combined or modified is not sufficient to establish prima facie obviousness. MPEP § 2143.01. Tobita does not teach that there is an interference signal or that it must be filtered out. Therefore, Tobita does not provide any suggestion or motivation to one skilled in the art to look to other circuits, such as Onishi, for interference signal filters.

Also, Onishi does not teach that there is a need to provide an automatic fine tuning circuit, such as Tobita's device, to control the frequency of the second local oscillator. In fact, Onishi teaches away from the voltage controlled oscillator ("VCO") of Tobita. The frequency of the second local oscillator of Onishi is disclosed to be a fixed frequency (column 7, line 60). Therefore, one of ordinary skill would not be motivated to modify Onishi by providing an automatic fine tuning circuit in which the second local oscillator is a VCO.

Additionally, the combination of the cited references does not teach all of the limitations of each claim. Claims 3 (now part of claim 1), 14-16, 19-23, 28 and 34 include the further limitation that the second mixer is an image rejection mixer. The Examiner takes Official Notice that an image rejection mixer is well-known and widely used in the art. However, there is no citation to any specific teaching in Tobita or Onishi that the combination of those references should be further combined with an image rejection mixer. Restating MPEP § 2143.01, the fact that the elements can be combined is not sufficient. In addition, it is not sufficient for the Examiner to simply state that an element is well-known. There is no teaching in

Tobita that there is an image frequency problem or that an image rejection mixer should be considered to eliminate an image disturbance. Furthermore, Onishi does not suggest the modification. The Onishi circuit eliminates image frequencies by the use of multiple filters throughout the circuit. There is no suggestion in the Onishi disclosure that further filtering by an image rejection mixer is required or desired. The Examiner has cited no reference that points to the use of an image rejection mixer in the manner claimed.

Applicant submits that the combination of Tobita and Onishi, as well as the further addition of an image rejection mixer, is not suggested by the references. Therefore, at the time the invention was made, one skilled in the art would not have been motivated to combine and modify the references to achieve Applicant's invention.

Onishi Reference

Claims 38, 41, 42, 44, 45, 48-55, 57, 90, 115 and 116 stand rejected under 35 U.S.C. § 103(a) as being unpatentable over Onishi.

The Examiner modifies Onishi by changing the second mixer to an image rejection mixer to provide grounds for rejecting claims 38 and 49. For the reasons stated above, Applicant traverses the statement that this is an obvious modification. Onishi discloses multiple filters which eliminate unwanted signals in the circuit. There is no suggestion or motivation in Onishi to change the second mixer to provide additional image signal rejection. This modification would have the effect of rendering other parts of the Onishi circuit redundant, which is contrary to the teaching of Onishi. Thus, by its own terms, the Onishi reference is not sufficient grounds for rendering the claims obvious under 35 U.S.C. § 103.

The Examiner also asserts that it is obvious to modify Onishi by changing the passband frequency of the band pass filter. Applicant submits that not only is there no suggestion or motivation for this modification, but that the disclosure teaches against this modification. There are multiple instances in the Onishi disclosure where it is stated that the pass band is 5 MHz at maximum. For example, "the band width [of the band pass filter] is 5 MHz so that interferences by adjacent channels in the following circuits are prevented to the utmost" (column 10, lines 57-59). Therefore, Onishi teaches away from widening the passband, as suggested by the

Examiner, because a larger pass band would allow adjacent channels to pass into the rest of the circuit.

The Examiner also takes Official Notice that an automatic gain control circuit is well known and widely used in the art. Again, Applicant submits that there is no suggestion or motivation for this modification to Onishi. Nothing in the disclosure suggests that the gain of the second amplifier could be, or should be, adjusted. Furthermore, there is no disclosure in Onishi which would motivate one skilled in the art to look for methods to control the gain of the second amplifier. Nor is there any teaching that the output signal strength from the second amplifier must be at a specified level.

Applicant submits that there is no suggestion or motivation to make any of the proposed modifications to Onishi that are suggested by the Examiner. Therefore, Onishi cannot be modified as proposed to render Applicant's invention obvious. Moreover, as discussed above with respect to the § 102(b) rejections, Onishi fails to disclose all of the limitations of the independent claims. Therefore, the Examiner's proposed modifications to Onishi do not disclose all the elements of dependent claims 38, 41, 42, 44, 45, 48-55, 57, 90, 115 and 116.

Claims indicated as allowable if rewritten

The Examiner indicates that claims 58-83 would be allowable if rewritten in independent form to include all of the limitations of the base claim and any intervening claims. Applicant submits that, in view of the amendments and arguments discussed above, claims 58-83 are in condition for allowance as depending from allowable base claims.

Applicant's attorney respectfully requests that the Examiner call him at the below listed number if the Examiner believes that such a discussion would be helpful in resolving any remaining problems.

Respectfully submitted,

FULBRIGHT & JAWORSKI, L.L.P.
Attorneys for Applicant

Date: March 12, 1997

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0394763
03/11/97

ATTORNEY DOCKET NO. 49581-P001US-964909 **PATENT APPLICATION**

IN THE U.S. PATENT AND TRADEMARK OFFICE

Inventor(s):	Robert Rudolf Rotzoll		
Serial No.:	08/426,080	Examiner:	Sherrie Hsia
Filing Date:	April 21, 1995	Group Art Unit:	2602
Title:	MONOLITHIC TELEVISION TUNER		

ASSISTANT COMMISSIONER
FOR PATENTS
Washington, D.C. 20231

TRANSMITTAL LETTER FOR AMENDMENT

Sir:

Transmitted herewith is/are the following in the above-identified application:

(X) Response/Amendment ( ) Petition to extend time to respond

( ) New fee as calculated below ( ) Supplemental Declaration

(X) No additional fee (Address envelope to "Box NON-FEE Amendment")

( ) Other: ____________________

At any time during the pendency of this application, please charge any fees required or credit any overpayment to Deposit Account 06-2380 pursuant to 37 CFR 1.25. Additionally, please charge any fees to Deposit Account 06-2380 under 37 CFR 1.19, 1.20 and 1.21.

I hereby certify that this correspondence is being deposited with the United States Postal Service as first class mail in an envelope addressed to: Assistant Commissioner for Patents, Washington, D.C. 20231

Date of Deposit: 3/12/97

Typed Name: Elva Wilson

Signature: Elva Wilson

Respectfully submitted,

FULBRIGHT & JAWORSKI L.L.P.

David H. Tannenbaum
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2200 Ross Avenue, Suite 2800
Dallas, TX 75201
Telephone: (214) 855-8333

Date: March 12, 1997

0403285

Notes:

TUNE 048331

TAB 7

49581-P001US-964909

PATENT

#11/D
mail
8-12-97

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

M. Graves
8-12-97

In re Application:	Robert Rudolf Rotzoll
Serial No.:	08/426,080
Filed:	April 21, 1995
Art Unit:	2602
Examiner:	Sherrie Hsia
For:	MONOLITHIC TELEVISION TUNER

8/15/97

Assistant Commissioner
for Patents
Washington, D.C. 20231

I hereby certify that this correspondence is, on the date shown below, being deposited with the United States Postal Service with sufficient postage as first class mail, in an envelope addressed to the Assistant Commissioner for Patents, Washington, DC 20231.

Date deposited: July 2, 1997

Signature: Maurie Katon

Printed Name: Maurie Katon

AMENDMENT

Dear Sir:

In response to the Office Action, mailed June 12, 1997, Applicant respectfully requests that the Examiner reconsider the application in view of the amendments and comments set forth below.

In the Claims

Please amend the claims as follows:

In claim 10, lines 1-2, replace "6" with --1-- after "Claim".

Please add the following new claims:

119 ~~120~~. The television receiver as set forth in claim 42 ~~43~~ wherein said input filter limits the frequencies passed to those at or below the highest channel in a desired frequency band.

120 ~~121~~. A television receiver comprising:

a receiver input for receiving an RF signal;

an input filter coupled to said receiver input and operating to remove all frequency components in said RF signal above an input cutoff frequency to produce a filtered RF signal;

a first reference signal having a first operating frequency;

a first mixer having a first input coupled to an output of said input filter and a second input coupled to said first reference signal;

a second reference signal having a second operating frequency; and

a second mixer having a first input coupled to an output of said first mixer and a second input coupled to said second reference signal, wherein said second mixer is an image rejection mixer.

D1

121 ~~122~~. The television receiver of claim 120 ~~121~~ wherein said first operating frequency is variable between a minimum frequency value and a maximum frequency value.

122 ~~123~~. The television receiver of claim 121 ~~122~~ wherein said minimum frequency value is greater than said input cutoff frequency.

123 ~~124~~. The television receiver of claim 120 ~~121~~ wherein said first mixer subtractively mixes said first reference ~~operating~~ signal and said filtered RF signal to thereby generate a first IF signal at said first mixer output.

124 ~~125~~. The television receiver of claim 120 ~~121~~ wherein said second mixer mixes a first IF signal from said first mixer and said second reference signal to generate a

second IF signal and wherein said second mixer rejects an image signal of said second reference signal from said second IF signal.

125 ~~126.~~ The television receiver of claim 120 ~~121~~ wherein said second mixer mixes a first IF signal from said first mixer and said second reference signal to generate a second IF signal and wherein said second mixer rejects an image signal of said first IF signal from said second IF signal.

126 ~~127.~~ The television receiver of claim 120 ~~121~~ wherein said first mixer generates a first IF signal and wherein a selected carrier frequency in said RF signal appears at a first predetermined frequency in said first IF signal.

127 ~~128.~~ The television receiver of claim 126 ~~127~~ wherein said second mixer generates a second IF signal and wherein said selected carrier frequency in said RF signal appears at a second predetermined frequency in said second IF signal.

128 ~~129.~~ The television receiver of claim 120 ~~121~~ wherein said first mixer and said second mixer are physically located on the same integrated circuit substrate.

Remarks

Claims 1, 2 and 4-119 are pending in the application. Applicant has added new claims 120-129. The Examiner has indicated that claims 35-119 are allowed. Claims 1-35 are either rejected or objected to by the Examiner. The outstanding issues in the instant application are:

-- Claims 1, 2, 4-9, 12, 13, 17 and 18 stand rejected under 35 U.S.C. § 103(a) as unpatentable over U.S. Patent No. 4,855,835 to Tobita ("Tobita") in view of U.S. Patent No. 3,839,676 to Linnecar ("Linnecar").

-- Claims 24-27 and 29 stand rejected under 35 U.S.C. § 103(a) as being unpatentable over Tobita in view of U.S. Patent No. 4,328,531 to Nagashima et al. ("Nagashima").

-- Claims 10, 11, 14-16, 19-23, 28 and 30-34 are objected to as being dependent upon a rejected base claim, but it is noted that these claims would be allowable if rewritten.

Applicant hereby traverses all outstanding rejections and objections and requests their reconsideration and withdrawal in view of the amendments and remarks contained herein.

<u>New claim 120</u>

Applicant has added new claim 120. This claim depends from base claim 43 which was indicated to be allowed in the current Office Action. Applicant submits that this new claim is patentable since it depends from an allowed base claim and thereby includes all limitations of the allowed claim.

<u>New claims 121-129</u>

Applicant has also added new claims 121 to 129. Independent claim 121 includes limitations which Applicant believes are indicated as patentably distinct over the prior art in view of the Examiner's rejections of and objections to claims 1-34. The Examiner has rejected claim 1 but indicated that dependent claim 10 would be allowable if rewritten. The significant limitation added by claim 10 is "an input filter which filters out all frequency components in said RF signal above an input cutoff frequency." Independent claim 24 has also been rejected. However, dependent claim 28 is indicated as allowable if rewritten. The unique limitation that is added in claim 28 is a second mixer which "rejects at least one image signal associated with said second operating frequency from said second IF signal."

New independent claim 121 includes both the limitations of an input filter and an image rejection mixer. Applicant submits that the prior art does not teach or suggest the combination of elements in claim 121 and, therefore, claim 121 is patentably novel and non-obvious. New claims 121-129 only comprise limitations which have been before the Examiner in prior pending claims and no new matter is added. Accordingly, dependent claims 122-129 are also patentable since they include the elements of base claim 121.

Rejection under Tobita/Linnecar combination

The Examiner has rejected claims 1, 2, 4-9, 12, 13, 17 and 18 under 35 U.S.C. § 103(a) as unpatentable over Tobita in view of Linnecar.

Claim 1 requires, among other things, a second mixer which is an image rejection mixer. Tobita discloses a second mixer, however, it does not teach the that the mixer is an image rejection mixer. Linnecar discloses the use of an image rejection mixer, however, it is not a mixer that is directly coupled to the output of a first mixer as required in claim 1. Therefore, the combination of Tobita and Linnecar do not teach or suggest all the limitations of claim 1 as required under M.P.E.P. § 2143.03.

Section 2143.01 of the M.P.E.P. states that some teaching, suggestion or motivation to combine or modify the references is required to establish obviousness. There is no teaching or suggestion in either Tobita or Linnecar to modify the second mixer of the Tobita circuit so that it is an image rejection mixer. The Tobita and Linnecar references can be combined. However, the resulting combination will merely be a more detailed version of the Linnecar circuit that does not teach Applicant's invention or the elements of claim 1. M.P.E.P. § 2143.01 states that the fact that references can be combined or modified is not sufficient to establish *prima facie* obviousness.

Linnecar teaches an apparatus for converting a satellite-broadcasted 12 GHz signal to a VHF or UHF signal for a television. The purpose of the Linnecar circuit is to allow for remote reception of a satellite signal which is then converted to a lower frequency for transmission over a feeder line to a demodulating circuit. In Linnecar, a 12 GHz signal is received by antenna 1 and then converted to a 40-70 MHz signal by mixer 11. Linnecar teaches that mixer 11 may be an image rejection mixer. The 40-70 MHz signal from mixer 11 is then provided to a power supply and demodulator circuit over feeder 13. Modulator/demodulator circuits 32 and 35 convert the 40-70 MHz signal to a VHF or UHF signal which is provided to input 37 of television 38. (Column 2, line 50 to column 3, line 16).

The receiver circuit used by television 38 is not taught in Linnecar. If the references were combined, the UHF or VHF band signal produced by Linnecar would

serve as the input to the Tobita receiver circuit and the Tobita circuit would be located within television 38. There is no suggestion or motivation in either reference to replace the second mixer of Tobita with the 12 GHz mixer of Linnecar. Also, there is no suggestion to move the 12 GHz conversion circuit and the power supply and demodulator circuit into television 38 or to combine it with the television's receiver. Instead, the television receiver and the 12 GHz conversion circuit are all shown to be discrete elements.

The combination of Tobita and Linnecar would effectively produce only a different version of Linnecar, not the circuit of claim 1. The combination would be missing the required element of an image rejection second mixer and, as a result, it would not teach or suggest all the limitations of claim 1. Therefore, Applicant submits that claim 1 is patentable over the combination of Tobita and Linnecar under 35 U.S.C. § 103(a). Accordingly, claim 1 and dependent claims 2 and 4-23 are also in condition for allowance.

Rejection under Tobita/Nagashima combination

The Examiner has rejected claims 24-27 and 29 under 35 U.S.C. § 103(a) as being unpatentable over Tobita in view of Nagashima.

Claim 24 requires a first mixer and a second mixer that are "physically located on the same integrated circuit substrate." Claim 29 requires that the RF and IF "mixing operations are accomplished on the same integrated circuit substrate."

Nagashima teaches a thick film multi-layer ceramic substrate having a plurality of interposed conducting layers and dielectric layers. The various conducting layers in Nagashima are designed to serve as connection electrodes for discrete circuit elements, such as transistors, diodes, coupling capacitors and coils (column 2, lines 41-44 and column 3, lines 65-67). As shown in Figure 3, the Nagashima invention is merely directed toward providing an improved circuit board for connecting the discrete tuner circuit elements. Nagashima does not teach or suggest physically incorporating the elements of a tuner circuit into an integrated circuit.

Nagashima does not show the elements of a mixer that can be "physically located on an integrated circuit substrate" or a second mixer "physically located on the same integrated substrate" as required by claim 24. Nagashima is also missing any teaching or suggestion of claim 29's method in which RF signal mixing and IF signal mixing are "accomplished on the same integrated circuit substrate." As a result, the Nagashima reference is missing elements of both claims 24 and 29.

M.P.E.P. § 2143.03 requires that all claim limitations must be taught or suggested by the prior art in order to support a *prima facie* case of obviousness. The Examiner admits that Tobita does not show mixers located on the same substrate. Since Nagashima also fails to provide mixers that are "physically located on an integrated circuit substrate," the combination of Tobita with Nagashima does not teach or suggest all of the claim limitations and, therefore, cannot render claims 24 and 29 obvious under M.P.E.P. § 2143.

In order to support a *prima facie* case of obviousness under M.P.E.P. § 2143.01, the prior art must provide some suggestion or motivation to modify the cited references. There is no suggestion or motivation in the Nagashima reference to physically locate mixers on an integrated circuit. Nagashima teaches that the power supply lines must be shielded between grounded conductors in order to prevent the local oscillator signal and intermediate frequency signal from leaking into the power supply lines (column 4, lines 4-27). It is well known that placing circuit elements, such as mixers, physically on an integrated circuit substrate causes signal coupling. The Nagashima reference fails to teach how to compensate for the additional coupling problems that would be caused by placing mixers on the substrate.

Figure 7 illustrates the coupling that occurs between the local oscillator signal and the power supply lines. The signal coupling shown in Figure 7 is for the Nagashima circuit which is comprised of discrete circuit elements. The disclosure admits that the circuit is "not always satisfactory at VHF band" (column 4, lines 64-68). Modification of the circuit to include mixers physically located on the integrated circuit substrate would likely cause even more signal coupling into the other circuit lines, thus defeating the purpose of Nagashima.

The combination of Tobita with Nagashima fails to teach or suggest all the limitations of claims 24 and 29. Also, there is no suggestion or motivation to modify the references to add the missing claim limitations. Therefore, the combination of Tobita and Nagashima does not satisfy the obviousness requirements of M.P.E.P. § 2143. Accordingly, Applicant submits that claims 24 and 29 are patentable under 35 U.S.C. § 103(a) over Tobita and Nagashima. Thus, dependent claims 25-27 are also patentable and in condition for allowance.

Claim objections

Claims 10, 11, 14-16, 19-23, 28 and 30-34 stand objected to as being dependent upon a rejected base claim, but the Examiner has indicated that they would be allowable if rewritten in independent form including all of the limitations of the base claim and any intervening claim.

In view of the above remarks, Applicant submits that base claims 1, 24 and 29 are allowable over the cited prior art. Therefore, the dependent claims, including claims 10, 11, 14-16, 19-23, 28 and 30-34, are also allowable over the prior art without further amendment.

For the above reasons, the Applicant submits that the claimed design is of patentable merit under 35 U.S.C. § 103. Accordingly, Applicant submits that this application is in full condition for allowance and asks that the Examiner pass this case to issue.

Applicant's attorney respectfully requests that the Examiner call him at the below listed number if the Examiner believes that such a discussion would be helpful in resolving any remaining problems.

Respectfully submitted,

FULBRIGHT & JAWORSKI, L.L.P.
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Date: 6/30/97

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0437022
6/26/97

CH0260288

OIPE 64477 JUL 07 1997 PATENT & TRADEMARK OFFICE

Attorney Docket No. 49581-P001US-964909

PATENT

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application:	Robert Rudolf Rotzoll
Serial No.:	08/426,080
Filed:	April 21, 1995
Art Unit:	08/426,080
Examiner:	Sherrie Hsia
For:	MONOLITHIC TELEVISION TUNER

RECEIVED AUG 12 97 GROUP 2500

Assistant Commissioner
for Patents
Washington, D.C. 20231

I hereby certify that this correspondence is, on the date shown below, being deposited with the United States Postal Service with sufficient postage as first class mail, in an envelope addressed to the Assistant Commissioner for Patents, Washington, DC 20231.

Date deposited: July 2, 1997

Signature: Maurie Katon

Printed Name: MAURIE KATON

Dear Sir:

TRANSMITTAL OF AMENDMENT

A. Transmitted herewith is an amendment in connection with this application.

STATUS

B. Applicant is

☐ a small entity. A verified statement

☐ is attached

☐ was already filed.

☒ other than a small entity.

EXTENSION OF TIME

C. ☐ Applicant petitions for an extension of time under 37 CFR 1.136 (Fees: 37 CFR 1.17(a)-(d)) for the total number of months checked below:

08/06/1997 SCORNICH 00000178 08426080
01 FC:102 60.00 OP
02 FC:103 198.00 OP

Extension (months)	Fee for other than small entity	Fee for small entity
☐ one month	$ 110.00	$ 55.00
☐ two months	$ 390.00	$195.00
☐ three months	$ 930.00	$465.00
☐ four months	$1,470.00	$735.00
		Fee $______

If additional extensions of time are required, please consider this a petition therefor.

TUNE 048347

☐ An extension for [ ] months has already been secured and the fee paid therefor of $___ is deducted from the total fee for the total months of extension now requested.

Extension fee due with this request $___

4. ☒ Applicant believes that no extension of term is required. However, this conditional petition is being made to provide for the possibility that applicant has inadvertently overlooked the need for a petition and fee for extension of time.

FEE FOR CLAIMS

5 The fee for claims (37 CFR 1.16(b)-(d)) has been calculated as shown below:

Col. 1	Col. 2	Col. 3	Small Entity		Other than a small entity	
CLAIMS REMAINING AFTER AMENDMENT	HIGHEST NO. PREVIOUSLY PAID FOR	PRESENT EXTRA	RATE	ADDIT. FEE	RATE	ADDIT. FEE
TOTAL* 128	MINUS** 119	9	X11=	$	X22=	$ 198.00
INDEP.* 11	MINUS*** 10	1	X40=	$	X80=	$ 80.00
☐ FIRST PRESENTATION OF MULTIPLE DEP. CLAIM			+130=	$	+260=	$
			TOTAL	$ [	TOTAL	$ 278.00

Additional claim fee due $ 278.00

TOTAL FEE PAYMENT

6 ☒ Attached is a check in the sum of $278.00.

☐ Please charge Deposit Account No. 06-2380 the sum of $______________.

A duplicate of this transmittal is attached.

FEE DEFICIENCY

7. ☒ If any additional extension or claim fee is required, please charge Deposit Account No. 06-2380.

Respectfully submitted,

FULBRIGHT & JAWORSKI L.L.P.
Attorneys for Applicant

By ______________________

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Date: July 2, 1997

0440824

Notes:

TUNE 048349

TAB 8

FA 15.5: A CMOS 1.8GHz Low-Phase-Noise Voltage-Controlled Oscillator with Prescaler

Jan Craninckx, Michiel Steyaert

Katholieke Universiteit Leuven, ESAT-MICAS, Heverlee, Belgium

The two HF components for a full CMOS 1.8GHz frequency synthesizing PLL, the VCO and prescaler, are realized. The low-phase-noise oscillator employs bondwires for the high-quality on-chip inductor. A special LC-tank design enables an even further reduction of the phase noise. The prescaler uses an enhanced ECL-like CMOS D-flipflop and has a fixed division ratio of 128. The VCO and prescaler are integrated in a standard 0.7µm CMOS process.

A general circuit diagram of an LC-tuned oscillator is shown in Figure 1. Its single-sided output spectral noise density (S) at an offset $\Delta\omega$ from the carrier is given by [1]:

$$S = kTR_{eff}[1 + A + 2Q](\omega_0/\Delta\omega)^2 df$$

The effective resistance R_{eff} is first-order equal to the total series resistance in the inductor and the capacitor. There are three contributions to the phase noise, reflected in the factor [1+A+2Q]. The terms 1 and A represent the noise of the parasitic resistances and of the amplifier, respectively. A is usually equal to or larger than 1. The term 2Q is present only if an active implementation of the inductor is used. Since this term is dominant, to achieve low power consumption, it is necessary to make a high-quality passive inductor on chip.

Passive inductors on chip are usually realized as spiral inductors. However, coupling with the substrate limits the maximum operating frequency and the series resistance usually is several ohms [2, 3]. In this design the parasitic inductance of a bondwire in an IC package is employed to realize a high-quality LC-tank. Gold bondwires have low resistance (<1Ω), and inductance value approximately 1nH per mm length [4,5]. Capacitance to the substrate is, apart from the bondpads, almost non-existent.

Since phase noise is proportional to the series resistance in the LC-loop, there is a technological limit on the minimum achievable noise. To go beyond that limit, an enhanced LC-tank can be employed. The basic concept behind this improvement is the creation of an effective signal in the LC-tank loop that is larger than the signal at the output ports of the LC-tank, which is limited by the 3V power supply. By making the signal in the LC-tank loop larger, the signal-to-noise ratio becomes larger as well. This is achieved by using multiple inductors and capacitors.

The LC-tank used in this design is shown in Figure 2. Four bondwires are used to make two inductors. The capacitors are split into two parts. The first one (C1a and C2a) is a Metal/Metal capacitor and is used to shift the DC-voltage at the internal node to ground. A metal/metal capacitor is used instead of a Poly/N+ one to limit the series resistance. The other capacitors (C1b and C2b) are junction capacitors that are used to tune the center frequency with the control voltage V_C.

The circuit diagram of the complete VCO is shown in Figure 3. Implementation in CMOS is possible since the amplifier circuit is f_T-independent. Indeed, the gate-source capacitance of M1 and M2 is placed in parallel with the LC-tank, so it can be neglected compared to the LC capacitance of a few pF. The only frequency-limiting factors are the poles formed by the drain resistance and capacitance. The oscillation frequency of 1.8GHz is chosen from application considerations. Simulations indicate, however, that operation at frequencies as high as 5GHz is possible.

The prescaler operates asynchronously and has a fixed division ratio of 128. It uses a cascade of 7 divide-by-2 circuits. The high frequency D-flipflop is shown in Figure 4. To achieve the 1.8GHz operation, the current sources normally present underneath transistors M1,M2 and M11,M12 are omitted. The drawback of this is the need for a correct input DC biasing. However, this is easily accomplished by the DC biasing of the VCO itself. The result is a 20% increase in operating speed. However, this D-flipflop is still the speed-limiting factor in this circuit.

A micrograph of the IC is shown in Figure 5. The die measures 1.4x 3.7mm². The on-chip bondwires can clearly be seen. The VCO output spectrum is shown in Figure 6. The center frequency is 1.76GHz, and the resolution bandwidth is 1kHz. The chip-to-chip variation in center frequency is less than 1%. This is because the bondwire inductance is mainly determined by the horizontal length of the bondwires, which is very well controlled if chip-to-chip bonds are performed. The single-sided spectral noise density at 10kHz from the carrier is -85dBc/Hz. The VCO tuning characteristic is shown in Figure 7. The tuning range is about 4.5% for tuning voltages up to 3V and almost 7% for tuning voltages up to 10V. The VCO operates on a single 3V power supply and consumes 8mA. The prescaler takes 7mA from a 4V power supply.

References

[1] Wang, Y.-T., A. A. Abidi, "CMOS Active Filter Design at Very High Frequencies," IEEE J. of Solid-State Circuits, vol. 25, no. 6, pp. 1562-1574, Dec., 1990.

[2] Nguyen, N. M., R. G. Meyer, "A 1.8-GHz Monolithic LC Voltage-Controlled Oscillator," ISSCC Digest of Technical Papers, pp. 158-159, Feb., 1992.

[3] Basedau, P., Q. Huang, "A 1-GHz, 1.5-V Monolithic LC Oscillator in 1-µm CMOS," Proc. of the 1994 European Solid-State Circuits Conference, pp. 172-175, Sept., 1994.

[4] Greenhouse, H. M., "Design of Planar Rectangular Microelectronic Inductors," IEEE Trans. on Parts, Hybrids and Packaging, vol. PHP-10, no. 2, pp. 101-109, June, 1974.

[5] Steyaert, M., J. Craninckx, "1.1-GHz Oscillator Using Bondwire Inductance," IEE Electronic Letters, vol. 30, no. 3, pp. 244-245, Feb. 3, 1994.

0-7803-2495-1 / 95 / $4.00 / © 1995 IEEE

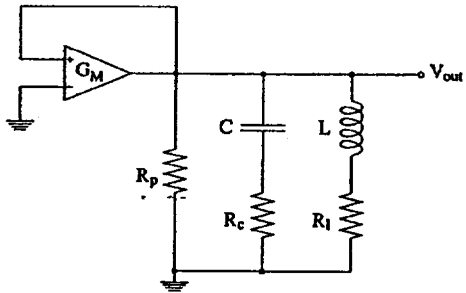


Figure 1: Basic LC-tuned oscillator circuit diagram.

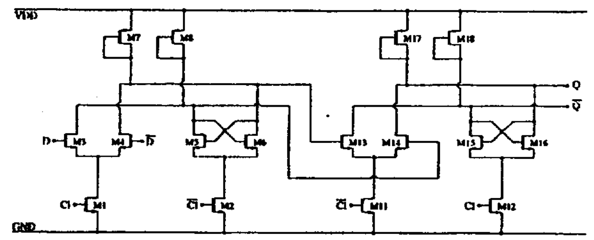


Figure 4: High-frequency D-flipflop.

Figure 5: See page 377.

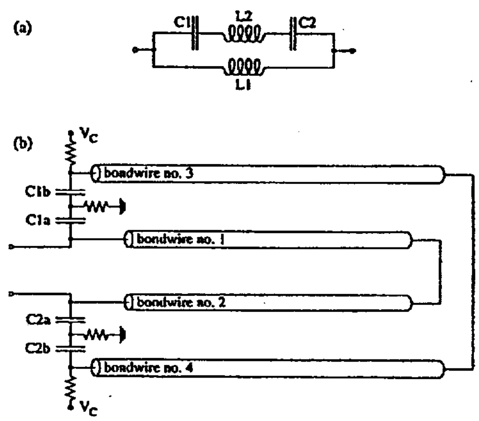


Figure 2: Enhanced LC-tank:
(a) general structure;
(b) realization with bondwires.

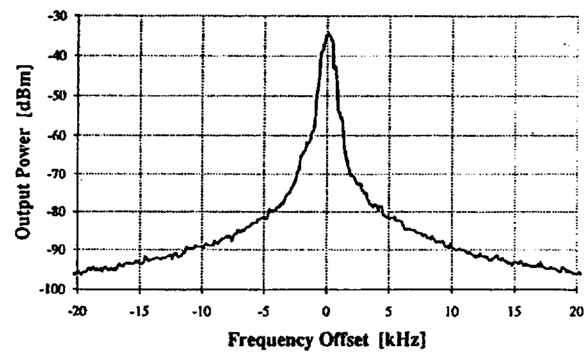


Figure 6: Output spectrum of the VCO.

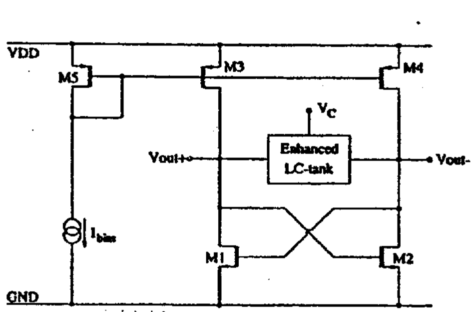


Figure 3: Circuit diagram of the VCO.

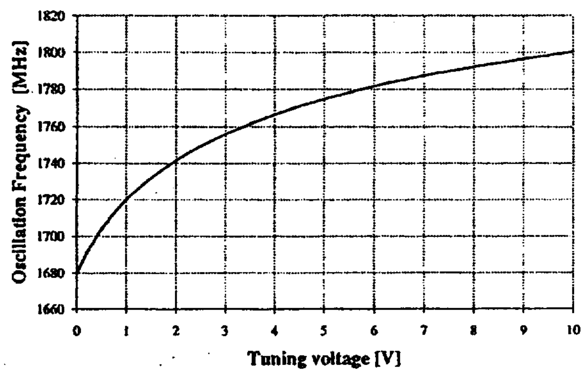


Figure 7: VCO tuning characteristic.

(Continued from page 267)

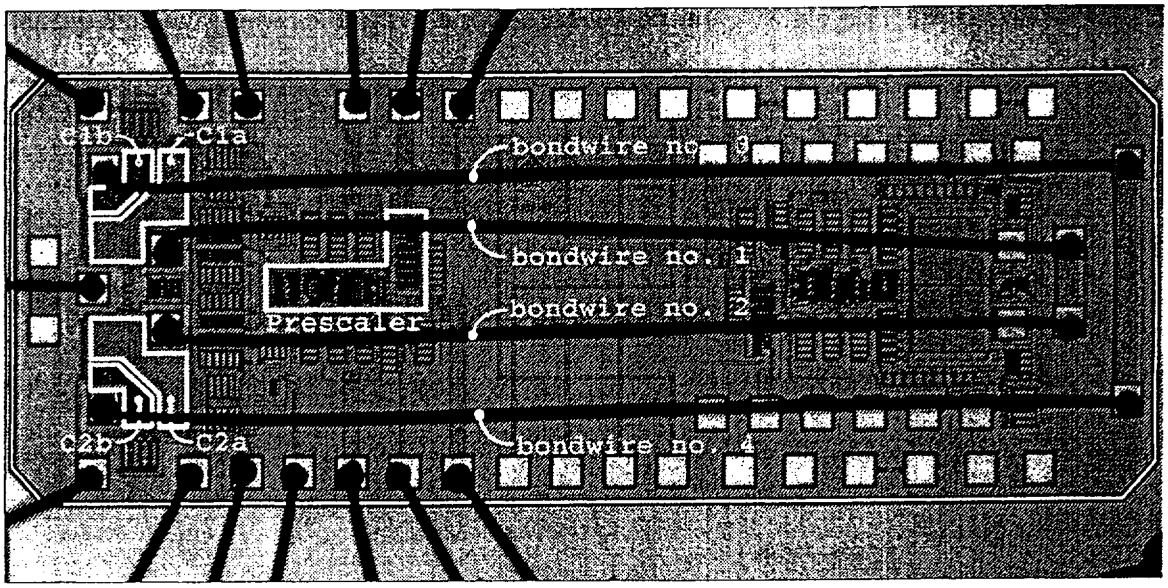


Figure 5: Chip micrograph.

FA 15.6: A 0.18µm CMOS Hot-Standby Phase-Locked Loop Using a Noise-Immune Adaptive-Gain Voltage-Controlled Oscillator (Continued from page 269)

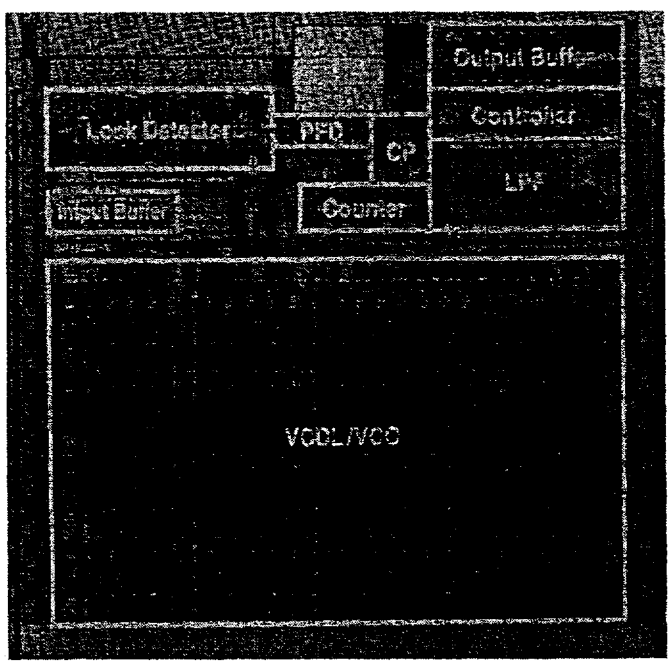


Figure 6: Chip micrograph.

FA 15.5: A CMOS 1.8GHz Low-Phase-Noise Voltage-Controlled Oscillator with Prescaler

Jan Craninckx, Michiel Steyaert

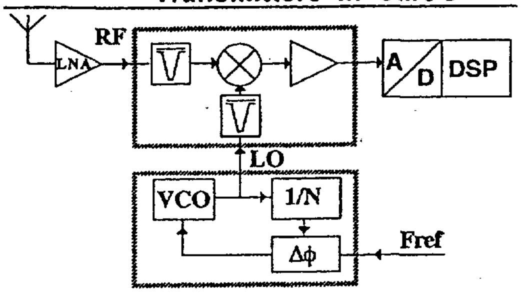
Katholieke Universiteit Leuven, ESAT-MICAS, Heverlee, Belgium

M. Steyaert

A 1.8GHz VCO using on-chip bondwire inductors demonstrates phase noise of -85dBc/Hz at 10kHz offset from the carrier. The chip in 0.7μm CMOS dissipates 24mW from a 3V supply.

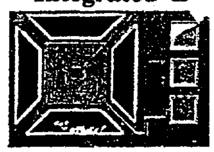
See Digest page 266.

ARTIC: Analog Receivers and Transmitters In CMOS



Inductor

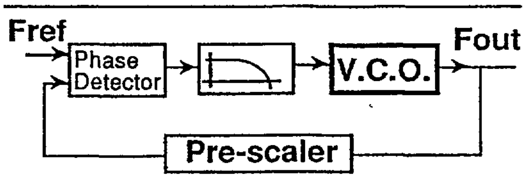
- No Active L ($\mathcal{L}(\Delta\omega) \sim 2Q!$)
- Integrated L



- High R (> 5 Ω)
- High Cp
- Low Q

☞ Etching ⇨ No standard CMOS

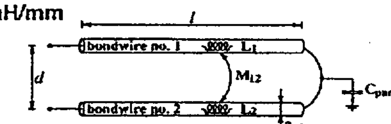
Frequency Synthesizer



- Standard CMOS
- Low Phase-noise
- No External Components

Bondwire Inductor

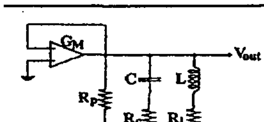
L ≈ 1 nH/mm



Advantages
- low R (< 1 Ω)
- no capacitance to substrate
- no extra processing cost

Disadvantages
- predictability
- reproducibility

LC-Tuned Oscillator



Active Gm

Active L

$$\mathcal{L}(\Delta\omega) = kT.R_{eff}[1 + A + 2Q].\left(\frac{\omega_0}{\Delta\omega}\right)^2$$

$$R_{eff} = R_l + R_c + \frac{1}{R_p(\omega_o C)^2}$$

Bondwire: Predictability

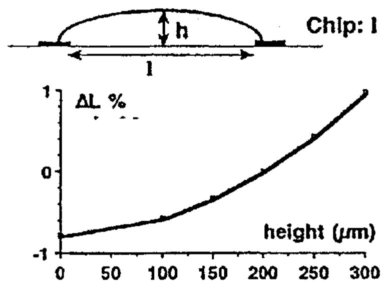
$$L_1 = L_2 = \frac{l}{5}\left[\ln\left(\frac{2l}{r}\right) - 0.75 + \frac{r}{l}\right]$$

$$M_{12} = \frac{l}{5}\left[\ln\left(\frac{l}{d} + \sqrt{1 + \left(\frac{l}{d}\right)^2}\right) - \sqrt{1 + \left(\frac{d}{l}\right)^2} + \frac{d}{l}\right]$$

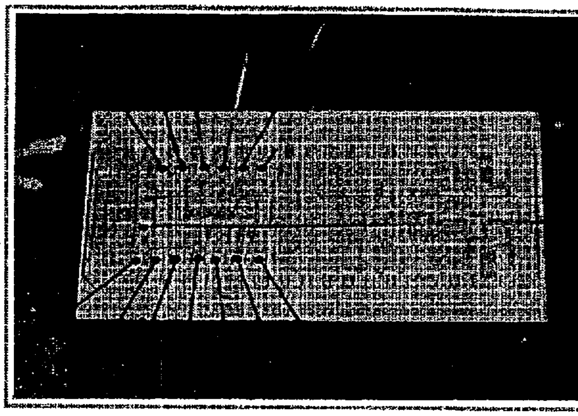
l = 4 mm
2r = 25 μm
d = 250 μm

L = 5.1 nH ± 5%

Bondwire: Reproducibility

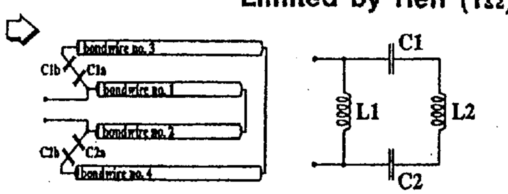


± 1%

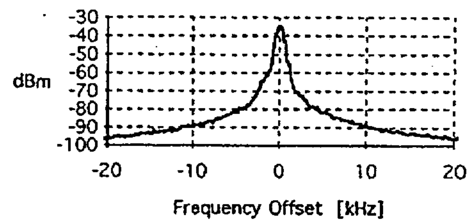


Enhanced LC-tank

$\pounds(\Delta\omega) = \frac{\text{Carrier}}{\text{Noise}}$ — Carrier: Limited by Vdd (3V); Noise: Limited by Reff (1Ω)

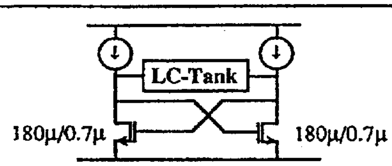


Measurement Results



fo = 1.76 GHz

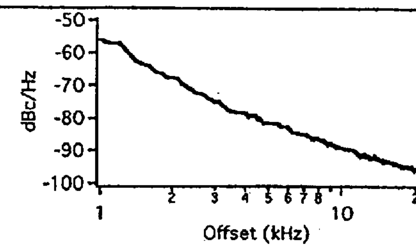
Oscillator Circuit



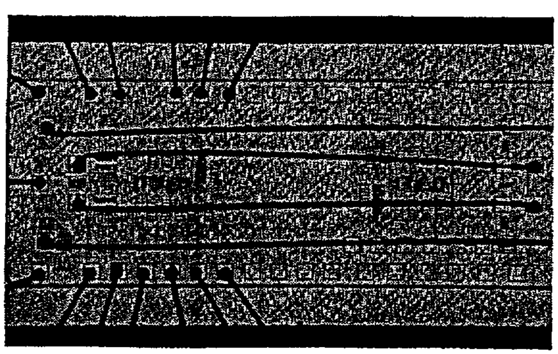
High Frequency operation:

- ft independent: Cgs // Ctank
- limited by drain R & C
- simulated up to 5GHz

Measured Phase-Noise

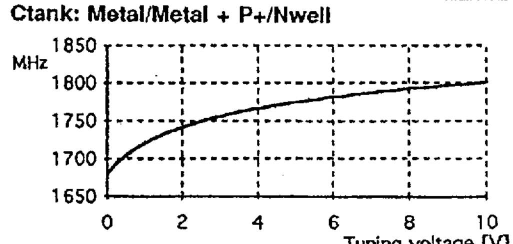


-85 dBc/Hz $\Delta\omega = 10$ KHz



Tuning Range

Ctank: Metal/Metal + P+/Nwell



5% ΔV = 3V

Continued on Page 362

Continued from Page 207

Conclusion

- Low power
  - PLL topology
  - ECL Stacking
  - BICMOS process to optimize blocks
- Integration
  - On chip VCO
  - On chip counters

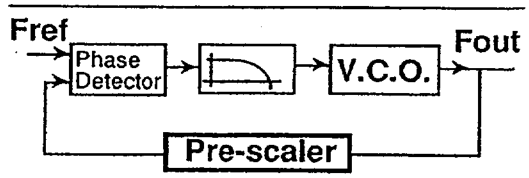
Limitations

- P+/Nwell: R => Tuning vs $\mathcal{L}$
  Lay-out P+ & Nwell contacts
- P+/Nwell biasing:
  Rbias => Modulation $\mathcal{L}$
- MOS in velocity saturation:
  white noise >> 4kT2/3gm

VCO Specifications

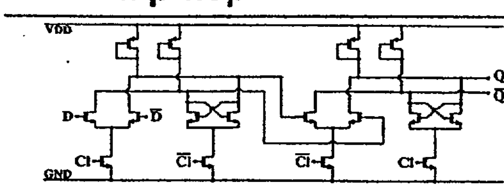
- 0.7µm Standard CMOS
- Bondwires: 2x 3.2 mm
  2x 3.6 mm
- fo 1.76 GHz
  Δfo < ±2%; ±1%
- Tuning 5% @3V
- $\mathcal{L}(\Delta 10\text{kHz})$ -85 dBc/Hz
- Power(3V) 24 mW

Frequency Synthesizer



- Standard 0.7µm CMOS
- 128 division

HF D-flip-flop



Advantage: speed + 20%

Disadvantage: DC biasing

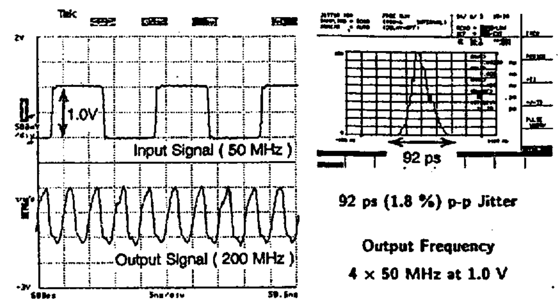
=> solved by VCO

Continued from Page 209

Pre-scaler Specifications

- 0.7μm Standard CMOS
- 7 devided-by-2: 128
  - 1° D-FF biased VCO
  - 2° Fully Dif D-FF
  - 3°-7° Dynamic CMOS
- Power 28 mW
- fin > 2 GHz

Measured Waveforms and Jitter



92 ps (1.8 %) p-p Jitter

Output Frequency

4 x 50 MHz at 1.0 V

Conclusions

Low Phase-Noise VCO in 0.7μm CMOS:

- Bondwire as inductor
- Standard CMOS
- $\pounds(\Delta 10kHz)$ -85 dBc/Hz
- 1.8GHz Pre-scaler

⇨ Single Chip Synthesizer

Specifications

Technology	0.18-μm CMOS
	Double Metal Layer
Number of Transistors	2,010
Voltage Supply	1.0 V
Power Dissipation	2 mW @ 200 MHz
Lock Time	< 500 ns @ 200 MHz
Lock Range	5 MHz to 200 MHz
Jitter	< 100 ps p-p @ 200 MHz
Die Area	0.48 mm × 0.45 mm

Summary

- Hot-Standby Architecture and Noise-Immune Adaptive-Gain VCO have been developed.
- Charge pump circuit and phase-frequency detector capable of operating at sub-1.0 volt have been developed.
- 2,010 Transistors have been integrated with 0.18-μm CMOS.
- 92-ps jitter and 200-MHz operation at 1.0 V have been achieved.

TAB 9

Future Directions in Silicon ICs for RF Personal Communications

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University of California, Berkeley

Abstract

This paper presents an overview of technical challenges in achieving higher integration levels, lower power dissipation, smaller form factor, and lower cost in portable battery-powered RF transceivers for personal communications applications. Specific emphasis is placed on silicon integrated circuits for transceivers in the 800MHz-2.5GHz range of frequencies.

Introduction

Digital radio personal communications devices utilizing the bands between 800MHz and 2.5GHz will play an increasingly important role in the overall communications infrastructure in the next decade. In addition, the bands above 2.5GHz, where large, relatively untapped blocks of spectrum are available, will receive increasing use as transceiver costs in this frequency range are brought down. Compared to other types of integrated circuits, the level of integration in the RF sections of such transceivers is still relatively low. Considerations of power dissipation, form factor, and cost dictate that the RF/IF portions of these devices evolve to higher levels of integration than is true at present.

In this paper, we attempt to identify some of the key barriers to realizing these higher levels of integration, and discuss several of the avenues currently being pursued for achieving that objective for portable personal digital RF communications devices such as cellular telephones, cordless telephones, wireless PBXs and wireless LANs utilizing the bands of frequencies between 800MHz and 2.5GHz. The emergence of established standards and the rapid growth of deployment make these very attractive potential applications for high-integration dedicated integrated circuits. Other important applications include the services to be offered by a whole spectrum of providers using both unlicensed bands as well as the licensed part of the new PCS band now being allocated around 1.8 GHz. A good overview of these applications is given in [1]

Typical Present RF Transceiver Implementation

The vast majority of currently-manufactured transceivers for the applications mentioned above utilize single- or dual-conversion configuration for the receive path. Baseband channel bandwidths range from 10kHz to 1-2MHz. Good examples of systems at the current state of the art are surveyed or described partially in [2][3][4][5][6]. A typical example of such a transceiver as might be used in a frequency-hopped wireless LAN application is shown in block diagram form in Fig. 1

The conventional architecture in Fig. 1. is not particularly amenable to higher levels of integration. Image rejection considerations usually dictate that the first intermediate frequency (IF) be on the order of 10% of the carrier frequency, with for example at least 70 MHz in 900 MHz receivers, and higher in higher frequency receivers. The use of complex signal representation at IF, as in image reject mixers, can improve image rejection by a considerable margin[47] and simplify the passive RF image reject filter, but the difficulty of phase and amplitude matching at IF usually limits the image rejection in such mixers to values on the order of 20dB, and as a result the IF frequencies must still be kept fairly high to preserve image rejection without requiring expensive and lossy ceramic RF filters. In many applications, two ceramic RF filters are required for adequate image rejection, one preceding and one following the LNA. Depending on the modulation scheme used, the range of frequencies over which the receiver must be tuned, and the amplitude of the near-carrier interfering signals, a second frequency conversion may be performed, translating the signal down to a second IF on the order of 10-20% of the first IF.

Most current implementations also utilize external varactor-tuned LC resonators to provide the tuning element of the voltage-controlled oscillator (VCO) or VCOs which, in conjunction with a crystal reference, provides for frequency synthesis of the local oscillator(LO). The relatively high Q required for these resonators stems from phase noise considerations, discussed later.

These particular aspects of receiver architecture have fundamental implications for receiver integration level. Unfortunately, the required high-Q, low-noise, low-distortion bandpass IF filter (70-100 MHz for 900 MHz receivers, for example) is well beyond the capabilities of current low-power integrated filter technologies. As a result, external high-Q passive filters are generally used, usually implemented with SAW filters, ceramic filters, or in some cases LC filters. Because of the frequencies involved and the package parasitics usually present, considerable power dis-

sipation is involved in taking IF signals off chip into these devices. Also, available on-chip spiral inductors in standard silicon technologies have Qs limited to 5-10 at the frequencies of interest, only adequate in some situations for implementation of the low-phase-noise VCO required in conventional synthesizer architectures.

In the transmit direction, the use of direct carrier modulation has become widespread[29], so that channel shaping filters can be implemented at baseband. In current practice the transmit power amplifier is usually implemented with GaAs discrete devices or simple ICs with a number of external inductors used for tuning and impedance matching. A narrowband external passive filter is usually required to limit transmit energy to the desired band.

The major challenge in RF transceiver design is to more effectively utilize scaled technologies to improve the integration level of RF transceivers, with resulting further improvements in power dissipation, form factor, and cost. Efforts are underway in industrial laboratories and universities around the world, taking various avenues toward this goal. The most promising approaches involve direct-conversion or low-IF receiver architectures that eliminate external IF filtering, new approaches to frequency synthesis that eliminate the need for external VCO resonators, and more effective utilization of on-chip spiral inductors available in near-standard IC technology to provide the tuning function essential to low-power realizations of RF functions. A hypothetical single-chip transceiver that might result from success in these areas is illustrated in block diagram form in Fig. 2

Direct-Conversion, Quasi-direct Conversion, and Low-IF Receiver Architectures

A promising direction in architectures for higher integration in RF transceivers is the use of zero IF, low-IF, or quasi-IF configurations in the receiver, following the pager model, and the use of direct modulation in the transmit path. These configurations have been investigated intensively for years (see for example [10][15]) but have made their way into practice in only a few specialized applications[13][14][15][11][12][16][18][19][17][20]. These configurations eliminate the external IF filtering function since the IF filter is replaced by two (I and Q) lowpass filters in the case of zero IF and quasi-IF receivers, or by a low-frequency, low-Q bandpass IF filter in the low-IF case.

The most severe problems in direct conversion receivers result from the fact that the baseband signal often contains low-frequency information that must be distinguished from DC and low-frequency errors that arise in the baseband signal path. One important error source is the device-mismatch-induced DC offset and 1/f noise of the signal path itself. For reasons of large-signal blocking performance, the gain of the LNA is usually restricted to the 20dB range, so that the wanted signal level reaching the mixer under weak signal conditions is on the order of 20300 microvolts in amplitude. The accumulated DC offset referred to the mixer output can easily be 10mV, 30-50 times larger than the signal. Another important contributor is LO leakage, resulting from the fact that since the LO is at the carrier frequency, any energy from it reaching the RF path demodulates to a DC offset. Because the effects of LO leakage can be a function of the impedance seen at the antenna, these DC offsets can vary with time in an unpredictable manner. The problem is more severe in frequency-hopping receivers because the carrier leakage is different at each hop frequency, giving a time variation to the DC offset that is induced. The cumulative effect of carrier feedthrough and DC offsets is to superimpose large, possibly time-varying additive errors on the

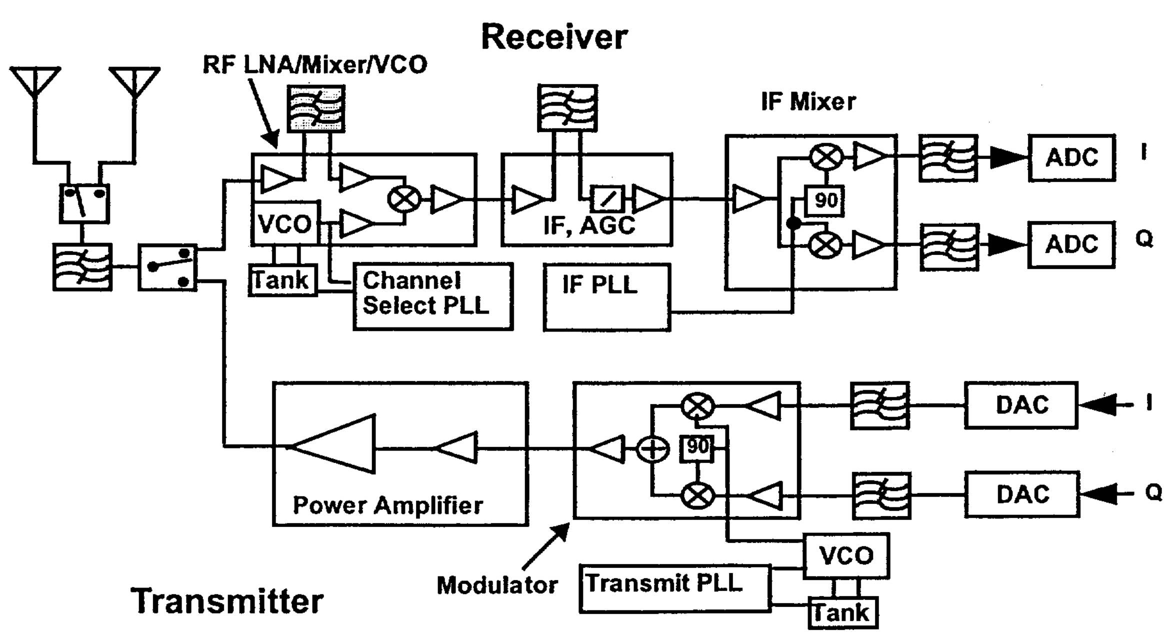

Figure 1. Block diagram of a typical multi-chip, multi-technology transceiver implementation

small wanted low-frequency AC signal in the baseband signal path.

Numerous approaches have been tried to attack these problems. Most have been attempted within the context of a more conventional, bipolar technology multi-chip receiver implementation using analog baseband filtering and demodulation. However, a closer coupling between the demodulation process and the RF and baseband analog signal path may well allow the separation of the DC offsets to be carried out using an adaptive approach that combines this function with carrier recovery, symbol timing recovery, automatic gain control (AGC), and data detection in a mixed analog-digital implementation. Most TDMA sytems, for example, utiliize a preamble in the frame structure which when demodulated to baseband has either zero or known DC content, allowing adaptive, frame-by-frame DC offset removal. The problem of 1/f noise can be attacked in a number of ways, one of which is to simply use correlated double sampling or chopper stabilization of the active elements in the baseband signal path. A/D conversion of the baseband signal at high resolution is a requirement for this approach.

Most of the benefits of homodyne receivers accrue if the IF is translated to a low but nonzero value instead of to DC as in homodyne receivers. The IF needs to be low enough that normal monolithic filtering techniques such as g_m/C continuous filters or switched-capacitor filters can be used. The advantage of this over homodyne receivers is that the problems of DC offset and 1/f noise are greatly reduced. However, a new problem of image rejection of the relatively close-in image frequency is introduced. This image energy must be eliminated through the use of an image-rejection mixer configuration following the LNA. Since this mixer will have to provide image rejection on the order of 60-70dB in some applications, phase shift accuracy and path matching accuracy within the mixer must be extremely precise. Progress has been made in this area in recent research [21].

Another important variation is the quasi-IF or "vestigial IF" approach in which the entire band of frequencies to be tuned by the receiver is translated down to IF in the first mixer, and then subsequently translated directly to baseband in the second mixer with little or no IF selective filtering. The channel-select filtering is done at baseband with a lowpass filter following translation to baseband. This technique has several important advantages. The first local oscillator can be implemented as a fixed frequency oscillator, making it easier to realize the required phase noise performance. The second LO, used to tune the desired channel, is at much lower frequency and its phase noise contributions, as well as the spurs associated with the narrow channel spacing and associated low comparison frequency, can be made much smaller. The carrier feedthrough problem is also eliminated. The technique eliminates the IF filter, but retains the image reject problem at RF and also many of the DC offset and drift problems of direct conversion receivers since adjacent channel blocking signals are carried to baseband and as a result most of the gain applied to the desired signal is done at baseband. The example receiver in Fig. 2 has this configuration.

Adaptation in Receiver Implementation

Conventional mixed-technology low-integration receivers have used RF/IF signal paths with relatively fixed functionality except for AGC and one or two other parameters. Higher integration implementations offer the possibility of much greater use of adaptive blocks. This capability may allow a signal transceiver to effect large savings in power, and to interface with more than one type of RF systems.

Power-Adaptive Transceivers

Most important power-dissipating elements in high-integration communications transceivers have a minimum dissipation requirement that is a function of the distance of the transceiver from the base station. In cellular phones, for example, the transmit power of the power amplifier is routinely varied adaptively depending on distance from the base station,both to save power and to reduce intrference to other users.

Improvements in overall average power dissipation could result from a wider application of the power adaptation concept. For example, the LNA power dissipation is dictated by the requirement for low noise figure and good input matching to accommodate the weakest signal that will be encountered far from the base station. When the signal is stronger, the LNA could be adaptively powered back to a lower power setting. In some current recievers a variation on this is now employed in which the LNA is simply bypassed and powered down for very strong signals. Similar adaptive power reductions may be possible in the mixer and baseband signal path as well. Another large dissipater is the synthesizer VCO, which must have very low phase noise for many applications. This noise is less important when the received RF signal is stronger, since the mixer-aliased VCO phase noise from the adjacent channel can be larger in absolute terms. As a result, power savings may be available by reducing VCO power and allowing larger phase noise. Similarly, the dynamic range required in the baseband signal path is greatest when the desired signal is weakest. Optimum distribution of gain and dynamic range through the baseband path is a function of botht he desired signal strength and the interfereing signal strength, both of which are easily detected in a digital implementation. For strong desired signal signal conditions, smaller filtering capacitors could be adaptively used

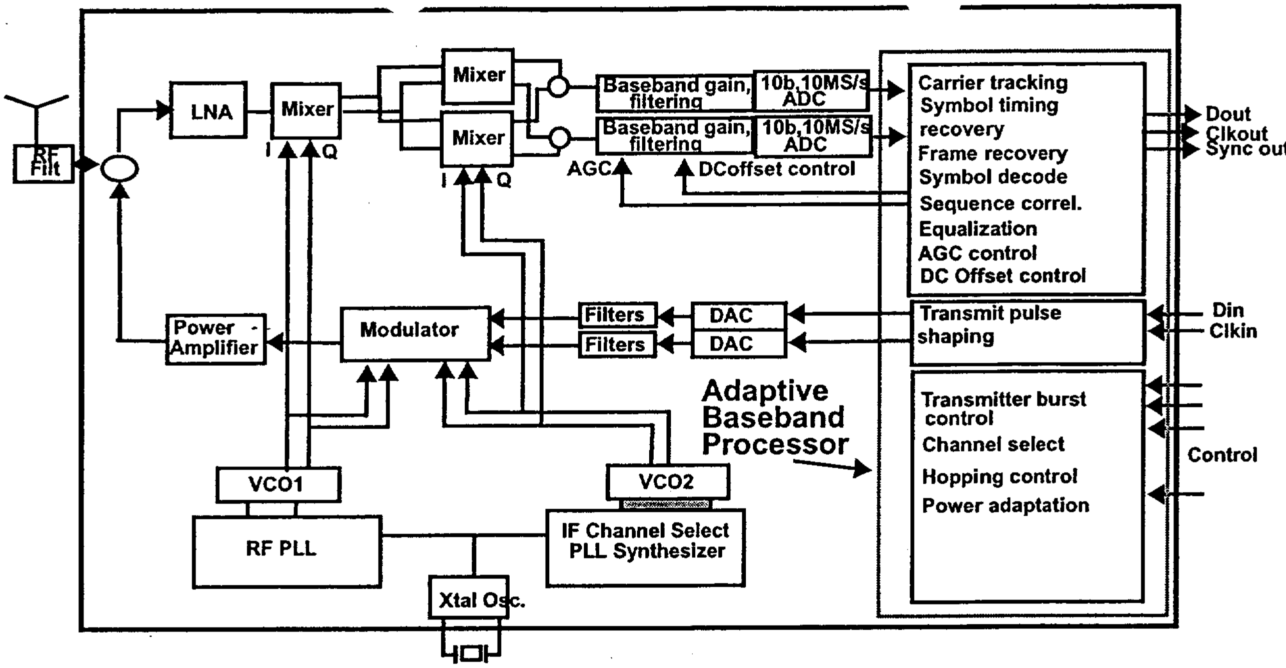

Figure 2. Block diagram of a possible future high-integration adaptive transceiver

in the path, allowing operation of the active devices at lower bias current for the same bandwidth.

Standard-Adaptive Transceivers

Great benefits could potentially accrue from a unification of approaches to data communications transceiver design. From an applications viewpoint, there is no fundamental reason that a single transceiver device could not provide the functionality of multiple communications standards at multiple frequencies. This might allow-for example, a single hand-held device to perform the functions of cellular phone and cordless phone compatibly with the varying standards for such service in Japan, North America, and Europe. Certainly multiple RF ports and antennas would be required for optimum performance at the different wavelengths, but following this a single transceiver should be able to adapt the different transmit power levels and modulation schemes. Central to achieving this would be an architecture in which a major portion of the IF (baseband in homodyne receivers) signal processing is performed in the digital domain, as well as carrier recovery, symbol timing recovery, equalization, DC offset control, power control, and so forth. A high-integration CMOS or BiCMOS implementation would be essential.

A secondary benefit would be a great reduction in overall engineering costs associated with development and productization of separate RF core functions for every standard and every type of device. The availability of a generic RF core function implemented in a VLSI digital compatible process such as scaled CMOS would be widely useful in implementing RF communications functions.

Circuit Approaches for Integrated Synthesizers

Most current transceiver implementations utilize external varactor-tuned resonators to implement the VCO function in one or two local oscillators that are slaved to a crystal reference through fixed or programmable divider chains in a phase-locked loop (PLL). Incorporation of these oscillators on-chip is an important goal in increasing receiver integration level and thereby reducing power. Often the divider is alternately switched between two adjacent divisor values at a high rate to achieve an effective interpolated value of division when very fine channel spacing is required(fractional-n synthesis). In the case of the US analog cellular standard, for example, the channel spacing is only 30kHz at 900 MHz, requiring the synthesizer to tune in very fine steps. This in turn requires the VCO frequency to be divided by a large integer, resulting in a low rate of comparison of current VCO phase with crystal oscillator phase. This in turn requires a low loop bandwidth.

Synthesizer PLL loop bandwidth is an important parameter in receiver design because of its influence on the local oscillator phase noise spectrum. The phase noise present in the local oscillator signal generated by the synthesizer contributes directly to phase noise on the IF or baseband signal after frequency conversion, and as a result directly degrades the effective signal-to-noise ratio (SNR) of phase-modulated signals. More importantly, LO phase noise mixes with adjacent channel signal energy, degrading overall receiver SNR and limiting receiver blocking performance. Finally, in the transmitter the phase noise of the LO contributes noise energy outside the band of the channel being transmitted. Spurious transmitted energy at adjacent channel frequencies must be closely controlled in most systems. With proper

PLL loop design, the phase noise of the synthesizer is dominated by the phase noise of the crystal reference for frequencies far below the PLL loop bandwidth, and by the inherent phase noise of the VCO itself for frequencies far beyond the PLL loop bandwidth.

In conventional synthesizers realizing fine channel spacing, the phase comparison frequency is low and loop bandwidth is low. As a result, synthesizer phase noise in the regions of interest is dominated by inherent VCO phase noise. For an LC oscillator, the ratio of the internally generated VCO phase noise power to the carrier power can be shown to be directly related to the amount of energy stored per cycle to the thermal energy kT. The energy stored is Q times the energy which must be supplied per cycle, and phase noise is directly related to the inverse of resonator Q, a fact predicted by a number of analyses of phase noise in oscillators.[31] Thus VCOs used to generate LO signals in phase-noise-critical applications almost always use some kind of external high-Q resonator.

A number of approaches show promise for realizing the VCO function on-chip. These include the use of on-chip spiral inductors[46], and the use of synthesizer configurations that allow wide PLL bandwidth so that the phase noise of the VCO is suppressed in the range of interest[27][28].

The use of on-chip inductors and varactors to implement the VCO has been demonstrated [14]using aluminum spiral inductors in standard bipolar technology. Because of the limited Q available, the phase noise achievable is not as low as required in some applications. Other alternatives with higher Q include the use of bond wire inductance [24][52]and the use of plated-up gold inductors over thick oxide [29]. Even with higher-Q inductors, the realization of a low-resistance, wide-range varactor tuning capacitance using standard IC technology is difficult. The simultaneous realization of high Q and tunability, together with either wide tuning range or high center frequency accuracy, is a very difficult task. The use of structures occurring in standard IC technology (such as bipolar base-emitter and base-collector capacitors) to perform the varactor function tends to introduce series resistance losses that reduce the Q to values below 10.

Ring-Oscillator-Based VCOs and Noise-Optimized Synthesizer Architectures

Ring oscillators are particularly attractive for LO generation because they inherently provide the quadrature clocks required for direct conversion and quasi-direct conversion receivers. Recent progress has been made in understanding the fundamental limits on noise in CMOS ring oscillator VCOs and the relationship between phase noise and power dissipation. In effect, for a given power dissipation a ring oscillator has a phase noise approximately equivalent to an LC oscillator with a Q of unity[27]. Each factor of 10 reduction in phase noise power with respect to the carrier requires a factor of 10 increase in power dissipation, all else being equal. The use of more advanced technology or lower supply voltages does not greatly alter this situation because of the fundamental processes involved. It appears that a phase noise level of about -106dBc per hertz should be achievable at 1MHz away from the carrier with about 50mW power dissipation in a 4-stage differential ring oscillator operating on a 3 volts supplyat 1.8Ghz. While this is adequate for some systems, it is inadequate for most digital radio communications applications.

The effects of close-in VCO phase noise can be minimized if PLL loop bandwidth, phase comparison frequency, and loop order can be kept high. A promising approach to achieving this is to use techniques of VCO phase interpolation in order to be able to make more frequent comparison of VCO phase with the instantaneous phase of the crystal reference. Discrete phase interpolation and noise-shaping M/N interpolation [49] are two examples of approaches to this goal. It appears likely that means can be found to increase the effective loop bandwidth by a large factor, thereby greatly reducing the effect of intrinsic VCO phase noise on the LO signal, in turn allowing the use of a ring oscillator VCO at least for some of the applications. In effect, this amounts to making the close-in VCO phase noise more dependent on the (very good) phase noise of the crystal reference and less dependent on the phase noise performance of the VCO. This approach has the additional advantage of preserving a wide loop bandwidth for frequency agility. Also, the short loop time constant potentially allows the powering down of the ring oscillator during inactive frame times in TDMA systems, resulting in power savings that offset the higher inherent active power of the ring oscillator.

The use of quasi-IF receiver architectures, in which the first VCO frequency is fixed, also relieves the phase noise problem because the comparison frequency and loop bandwidth can be kept much higher in a fixed-frequency first VCO. The second VCO/synthesizer performs the tuning function, but its impact on receiver phase noise is smaller since it operates at a much lower frequency.

Direct Digital Synthesis of the LO Signal.

Direct digital synthesis of the LO signal using a DAC, ROM, and phase-accumulation synthesizer is a very attractive alternative due to the excellent frequency agility achieved. Considerationsof power dissipation and technology speed capability limit such waveform synthesis to about 100MHz and below for typical current technologies. However, the synthesized waveform can be used as a frequency offset added to a carrier generated with a fixed-frequency VCO and PLL[18][48]. Because the generated frequency is

a fixed integer multiple of the crystal reference, the bandwidth of this loop can be kept high, allowing low phase noise in this fixed reference signal. This small difference frequency is used to translate the fixed carrier from the first PLL to the final carrier frequency using a quadrature modulator. It appears that this will be a very effective solution for at least the subset of applications where modest phase noise and modest spurious output component requirements can be tolerated. A similar approach using a secondary PLL to generate the offset frequency has recently been proposed[29].

Low-Power Baseband Signal Conversion and Processing

Depending on the type of transceiver, baseband operations of IF filtering, equalization, timing recovery, symbol constellation decoding, signal correlation, symbol generation, quadrature modulation, frequency synthesis and so forth are required. A major body of current research is aimed at performing more of these functions in the digital domain than is currently the case, with resultant improvement performance, adaptability, and manufacturability. The principal trade-offs are the incurred penalty in power dissipation and die area of the digital implementation, and the cost and die area of the required A/D converter. Rapid progress has been made in the implementation of these types of functions in VLSI CMOS [38][37][39] [19].

Many benefits accrue in pushing baseband signal processing into the digital domain, particularly for multistandard adaptive transceivers. For direct conversion receivers, the composite baseband signal contains all the large adjacent-channel blocking signals, and as a result an all-digital implementation of the baseband signal processing would require two A/D converters of greater than 80dB dynamic range and 20MHz effective sampling rate. Some combination of analog and digital filtering will be optimum. For at least the higher-frequency portions of this set of applications, low-power, high-speed approaches such as pipelining will be required. Finding techniques for reducing the power dissipation of these A/D converters is a key goal. Current state of the art for this class of converters is about 1mW/ MHz of sample rate at 10 bits [35].

Technologies for High-Integration RF Transceivers

Current transceiver implementations usually use a mix of technologies, with GaAs for the power amplifier and perhaps for the LNA, bipolar or BiCMOS for the mixer and IF functions, and CMOS for the baseband processing. High-integration implementations will require use of a single technology for most of the functions. GaAs will continue to play a very important role at the higher frequencies, but it appears likely that high-integration all-silicon solutions will evolve at the lower end of the spectrum.

Alternative technologies for a transceiver at the integration level of Fig 2 are BiCMOS and CMOS. Bipolar and BiCMOS solutions are attractive because of the inherent capability of bipolar transistors to provide high g_m at low current, and because of the well-developed family of circuit techniques for RF design using bipolar technology[23]. There is also considerable interest in utilizing CMOS for high-integration transceivers, particularly within the university community. Because of the potentially lower cost of a CMOS implementation, efforts to overcome the poorer characteristics of CMOS for RF by utilizing alternative receiver architectures, taking advantage of the high f_{max} of the NMOS device, and using more adaptation in the receiver may pay large dividends. The continued scaling of CMOS technology, with 0.1 micron devices with f_ts of near 100GHz recently demonstrated,[42][43] should eventually allow this approach.

Silicon-germanium technologies now evolving have the potential to provide bipolar devices with substantially larger f_t and f_{max} than the best current bipolar technologies[44]. The impact of this development is likely to be felt mostly at the high end of the frequency spectrum under discussion here.

Compatible High-Q Inductors and Resonators in Silicon

For reasons explained earlier, on-chip inductors are essential to low-power RF design. Present practice makes extensive use of wirebond inductance and spiral inductors in silicon in products currently in production. Improvement of the implementation of and the modeling of integrated inductors is a key goal.

Bond Wire Inductors

Bond wires provide inductance on the order of 1-4nH, depending on length, at 2 GHz, with Q on the order of 50. Higher inductance is realizable with unusual pad placements.[24][52] While pad-pad jump bonds are possible, they are not compatible with most automated bonding equipment. Inductor tolerance is a function of die attach, bonder mechanical accuracy, and wire diameter, and is in the +/- 20% range at present.Adjustments to mechanical assembly procedures could improve this significantly. Matching of bond wires on one side of the die to the same post is also better than 20%. Mutual coupling effects of adjacent wires must be accounted for through electromagnetic analysis of the configuration with a commercial package.

Creative use of bond wire inductance in SO and SSO packages with (perhaps) custom lead frames is a highly promising approach to implementing matching and tuning inductors in 1 and 2 GHz LNAs, VCOs, and power amplifiers. These elements are already widely used to perform part of the impedance matching function at LNA inputs and in

power amplifiers.The tolerance is a major problem, since there is no good high-Q variable capacitive element available.

Spiral Inductors on Silicon

Extensive work has been carried out on spiral inductors on silicon[45][46]. These devices can provide implement inductors up to about 10nH range in reasonable area, with Q limited to about 5 at 1 GHz and 10 at 2GHz by metal series resistance for standard technologies. Self-resonance due to the large parasitic capacitance to the substrate is a substantial problem, and drops to about 2GHz for a 10nH inductor in typical technology. Since the inductor is usually being used to match impedence or tune a gate or base diffusion capacitance, the parasitic capactiance can usually be incorporated in the design process as long as the self-resonant frequency is far above the frequency of interest.

Two approaches have been described in recent work for improving the performance of spiral inductors. In one approach, [25] a pit is etched under the spiral to remove the substrate capacitance and resulting self-resonance. The resulting structures have moderate Q but good self-resonance characteristics. Another approach is the deposition of thick oxide following normal IC fabrication, and the deposition of highly conductive interconnect metallization such as gold to form a high-Q spiral with reasonable self-resonance behavior. [29] This approach has high promise because the subsequent processing is relatively non-invasive to the underlying silicon and involves only low-temperature deposition and masking steps.

Package and Substrate Modeling

Perhaps the greatest single barrier to higher integration in RF transceivers is undesired interactions through substrate and package coupling. The problem can be addressed through a number of architectural and circuit approaches, but a critical missing link is a substrate and package modeling and simulation methodology to allow accurate prediction of these coupling effects prior to fabrication. Rapid progress is being made in this area [41][33][50].

Summary

Prospects for continued progress in high-integration, low-cost RF transceivers is excellent. A key requirement for progress is close collaboration between transceiver and system designers and architects, RF and digital circuit designers, and device and package modeling engineers, so that opportunities for innovation with new architectural approaches can be identified and exploited.

Acknowledgements

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TAB 10

Noise in Distributed MESFET Preamplifiers

Al P. Freundorfer, *Member, IEEE*, and Thé Linh Nguyen

***Abstract*—The theory of noise in a distributed MESFET preamplifier is developed. From this, it is shown that the equivalent input noise current density of a distributed preamplifier of an optical receiver can be improved by using large gate line matching impedance and appropriate scaling of the MESFET width. A front-end tuning circuit was designed using filter theory to further improve the noise performance of the preamplifier. A monolithic GaAs MESFET distributed preamplifier was fabricated with on chip front-end tuning components. Using a 35 μm InGaAs p-i-n photodiode, the preamplifier was shown to have an equivalent input noise current density of 8 pA/$\sqrt{\text{Hz}}$ and an 8 GHz bandwidth. To date, this is the best known result for a 0.8 μm GaAs MESFET process.**

I. Introduction

As the bit rate of optical communications systems increases into the 10 Gb/s region and beyond, there is interest in the development of large bandwidth low noise receivers [1]–[23]. These receivers include the high input impedance (HZ) [1]–[10], the transimpedance (TZ) [11]–[18], and the distributed (DA) and/or traveling wave (TW) preamplifiers [19]–[23] shown in Fig. 1(a)–(c).

Tuning networks are often used to reduce high frequency noise in the HZ preamplifier [2], [3], and lossy matching networks are frequently used with HZ to get good bandwidth and noise performance [1], [5], [6]. Current research in optical front-end receiver design is improving the noise-bandwidth performance through the use of improved device technology as well as circuit topologies [24]. The results of these preamplifiers [1]–[20] are plotted in Fig. 1(d). The noise-bandwidth performance is significantly improved by using superior circuit topologies from transimpedance to high impedance and distributed design, and by using superior device technology from MESFET to high electron mobility transfer (HEMT).

It was shown that distributed amplifiers have greater bandwidths than the feedback amplifiers for device technology with the same f_T and $f_{\max}$ [24]. The DA preamplifier topology also has a noise performance similar to the HZ preamplifier. The first demonstrated DA preamplifier had a bandwidth of 23 GHz with an input noise of greater than 20 pA/$\sqrt{\text{Hz}}$ [19] and was fabricated using a MESFET process that utilized a gate length of 0.3 μm. Recently, the distributed preamplifier was demonstrated using a HEMT process which had an improved noise performance of 15 pA/$\sqrt{\text{Hz}}$ over the operating frequency band of 2–18 GHz [23]. It was first shown that the gate line can be increased to 100 Ω to obtain better noise performance (8 pA/$\sqrt{\text{Hz}}$) while maintaining sufficient bandwidth for system specification [20]. A HEMT TW preamplifier with input noise current density of 20 pA/$\sqrt{\text{Hz}}$ and a 30 GHz bandwidth was demonstrated for the purpose of 40 Gb/s lightwave transmission [21].

In this paper, noise analysis will be given and methods for optimizing the performance of the DA preamplifier will be discussed. The theory of DA preamplifier will be derived in which the gain and the total equivalent input noise current density will be obtained. It will then be shown that this noise can be reduced by increasing the gate line impedance and using the optimal gate width. A new method for front-end tuning will be given for the DA preamplifier. Finally, to demonstrate these principles, a single photodiode distributed preamplifier was constructed using a 0.8 μm GaAs MESFET process available at Nortel (NT) and a 35 μm InGaAs p-i-n photodiode from NT. This distributed preamplifier was shown to have a bandwidth of 8 GHz. It also has an equivalent input noise current density of 8 pA/$\sqrt{\text{Hz}}$, which gives a predicted receiver sensitivity of −21 dBm for a 10 Gb/s intensity modulated direct detection (IM/DD) system [25]. To date, this is the best known result for a 0.8 μm GaAs MESFET process.

II. Theory

A. DA Gain Analysis

The simplified circuit model of a MESFET used in the gain analysis of the DA is shown in Fig. 2(a). The capacitance c_{gd} was absorbed in c_{ds} and c_{gd} using the Miller theorem. The Miller effect is small for the DA structure since the gain across the transistor is small.

The circuit representation of the DA shown in Fig. 1(c) is redrawn with the model in Fig. 2(b). The thermal noise models for the matched termination resistors $Z_{\pi g}$ and $Z_{\pi d}$ are given by $i_{Z_{\pi g}}$ and $i_{Z_{\pi d}}$, respectively. Also note that the noise of the transistor is shown only for the rth stage of the DA preamplifier. Since interest is only in the gain analysis, noise sources will not be used until our noise analysis in Section II-B of the theory. The two artificial transmission lines consist of the lumped inductors and the capacitive loads of the gate and drain, forming the gate and drain lines, respectively. The two lines are coupled through the transconductance of the transistors. A wave generated at the input of the gate line propagates with a delay of $\phi_g = \beta_g d = \omega\sqrt{L_g C_{gs}}$ and gets dissipated in the gate termination $Z_{\pi g} = \sqrt{L_g/C_{gs}}$. The voltage that appears across each gate produces a drain current through the transistor's transconductance. The drain current then propagates in two directions, one toward the

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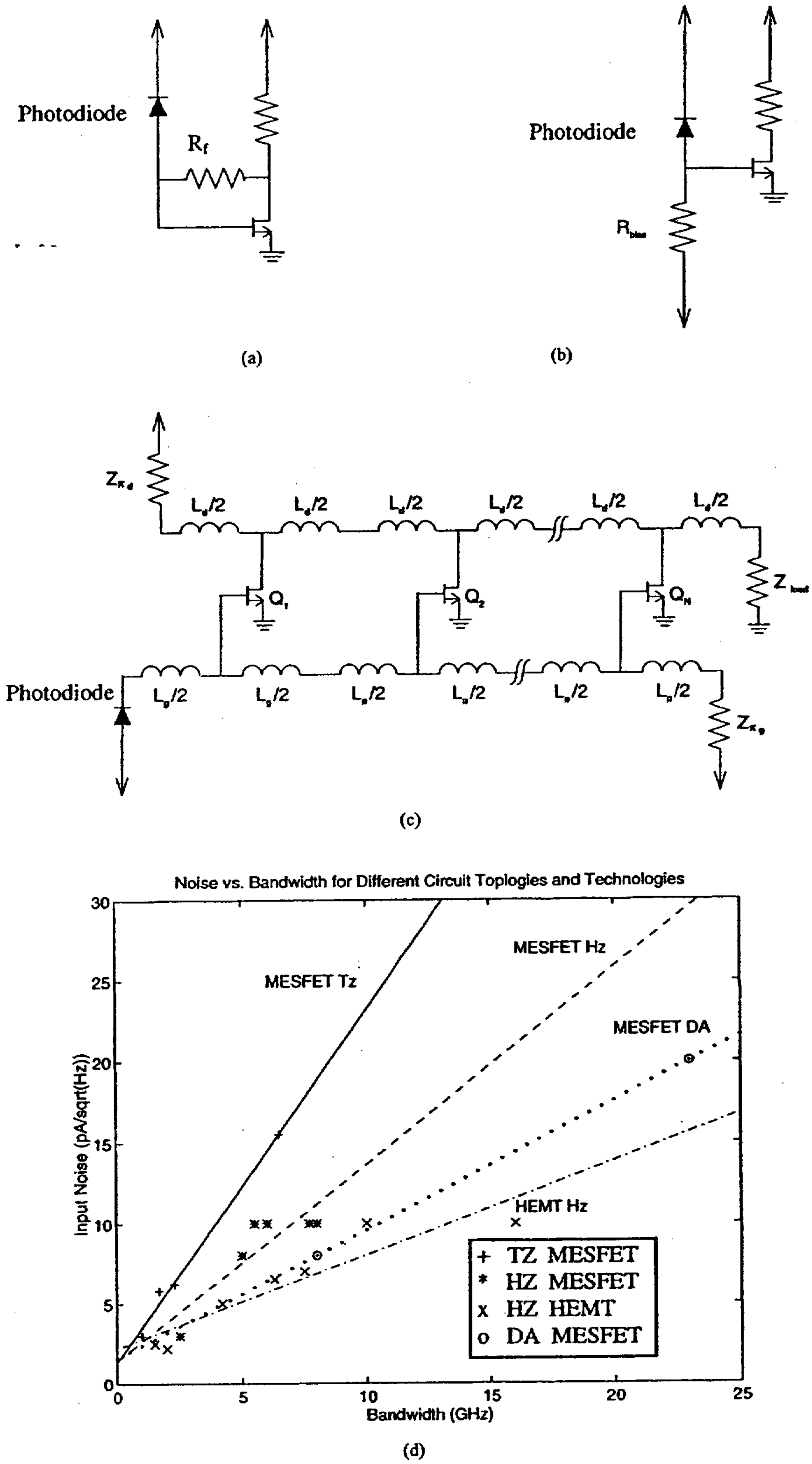
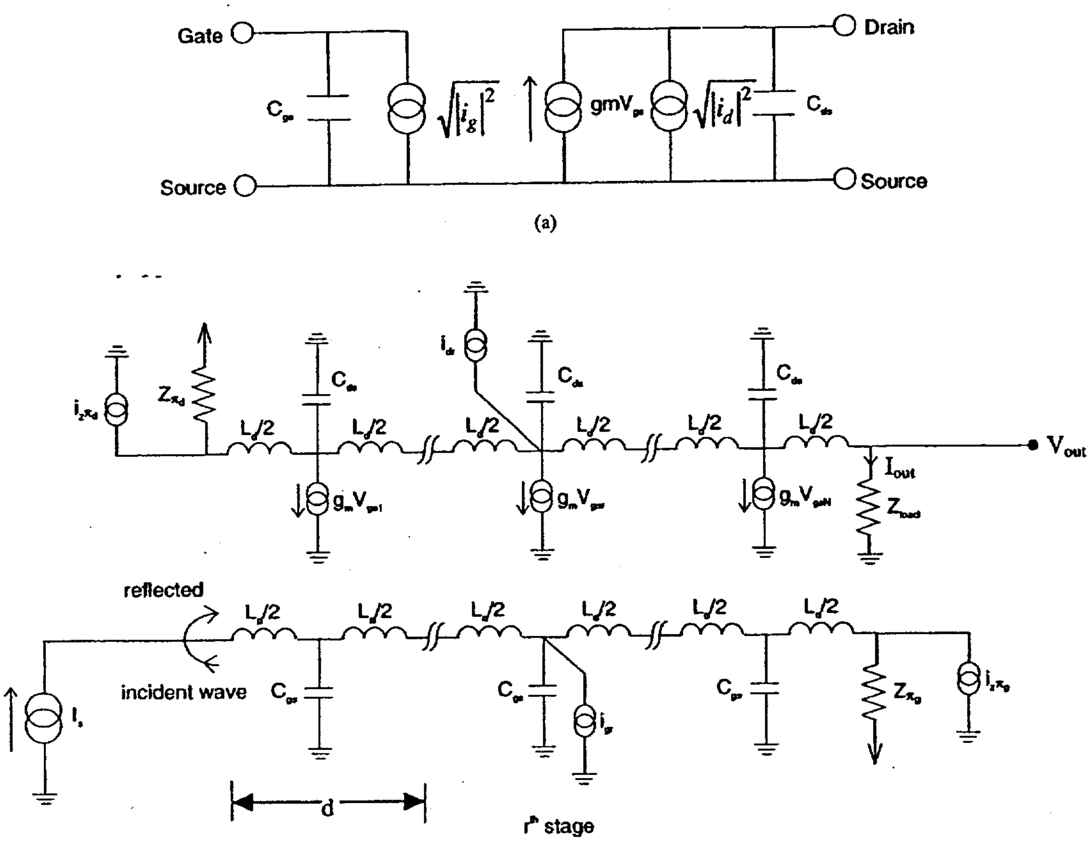


Fig. 1. (a)–(c) Preamplifier circuit topologies, (d) noise-bandwidth performance of various circuit topologies and technologies.

drain termination, and the other toward the load, with a propagation delay $\phi_d = \beta_d d = \omega\sqrt{L_d C_{ds}}$. This current wave eventually gets dissipated in both the drain termination, $Z_{\pi d} = \sqrt{L_d/C_{ds}}$, and the output load. The total gain at the output load is called the forward gain, and the total gain at the drain termination is called the reverse gain.



(a)

(b)

Fig. 2. (a) Simplified model of a MESFET with its associated gate and drain noise. (b) Small signal circuit representation of DA preamplifier.

At first we will assume different propagation delays, i.e., $\phi_d \neq \phi_g$, and then we will impose phase match where the propagation delays are the same. Under phase matched condition, the forward gain becomes constant as a function of frequency. In pulse shaping DA's [26], $\phi_d \neq \phi_g$ and thus the more general expressions are applicable.

1) Forward Transimpedance Gain: The forward transimpedance gain of the DA preamplifier is defined as the ratio between the voltage that appears at the output load and the input current. Here one assumes that the photodiode is ideal to simplify the analysis. It will be shown later that the photodiode capacitance will be absorbed in the front-end tuning network. I_{out} is defined as the total current delivered to the load by a DA with N stages, and the following expression can be obtained:

$$\begin{aligned} I_{\text{out}} = &-\tfrac{1}{2}(I_1 \exp[-j(N-\tfrac{1}{2})\phi_d] + I_2 \exp[-j(N-\tfrac{3}{2})\phi_d] \\ &+ \cdots + I_N \exp(-j\tfrac{1}{2}\phi_d)) \end{aligned} \tag{1}$$

where

$$\begin{aligned} I_i &= \text{drain current at } i\text{th stage, } i = 1, 2, \cdots, N \\ &= g_m V_{gs_i} \end{aligned} \tag{2}$$

and

$$V_{gs_i} = V_{\text{in}} \exp[-j(i-\tfrac{1}{2})\phi_g]. \tag{3}$$

Substituting (2) and (3) into (1)

$$\begin{aligned} I_{\text{out}} &= \tfrac{1}{2} g_m V_{\text{in}}([\exp -j((N-\tfrac{1}{2})\phi_d + \tfrac{1}{2}\phi_g)] \\ &\quad + [\exp -j((N-\tfrac{3}{2})\phi_d + \tfrac{3}{2}\phi_g)] \\ &\quad + \cdots + \exp[-j(\tfrac{1}{2}\phi_d + (N-\tfrac{1}{2})\phi_g)]) \\ &= -\tfrac{1}{2} g_m V_{\text{in}} \exp -j((N-\tfrac{1}{2})\phi_d + \tfrac{1}{2}\phi_g) \\ &\quad \cdot \sum_{n=0}^{N-1} e^{-jn(\phi_g - \phi_d)} \end{aligned} \tag{4}$$

using the identity $\Sigma_{n=0}^{N-1} a^n = -(1 - a^N/1 - a)$ one obtains

$$\begin{aligned} I_{\text{out}} = &-\frac{1}{2} g_m V_{\text{in}} \exp\left[-j\left(\left(N-\frac{1}{2}\right)\phi_d + \frac{1}{2}\phi_g\right)\right] \\ &\cdot \frac{\exp\left[-j\frac{N}{2}(\phi_g - \phi_d)\right]}{\exp\left[-j\frac{1}{2}(\phi_g - \phi_d)\right]} \frac{\sin\frac{N}{2}(\phi_g - \phi_d)}{\sin\frac{1}{2}(\phi_g - \phi_d)}. \end{aligned} \tag{5}$$

Under the condition of phase matching, $\phi_g = \phi_d = \phi$, (5) becomes

$$I_{\text{out}} = -\tfrac{1}{2} g_m V_{\text{in}} N e^{-jN\phi}. \tag{6}$$

And since

$$V_{\text{in}} = I_S Z_{\pi g} \tag{7}$$

$$I_{\text{out}} = -\tfrac{1}{2} g_m N I_S Z_{\pi g} e^{-jN\phi}. \tag{8}$$

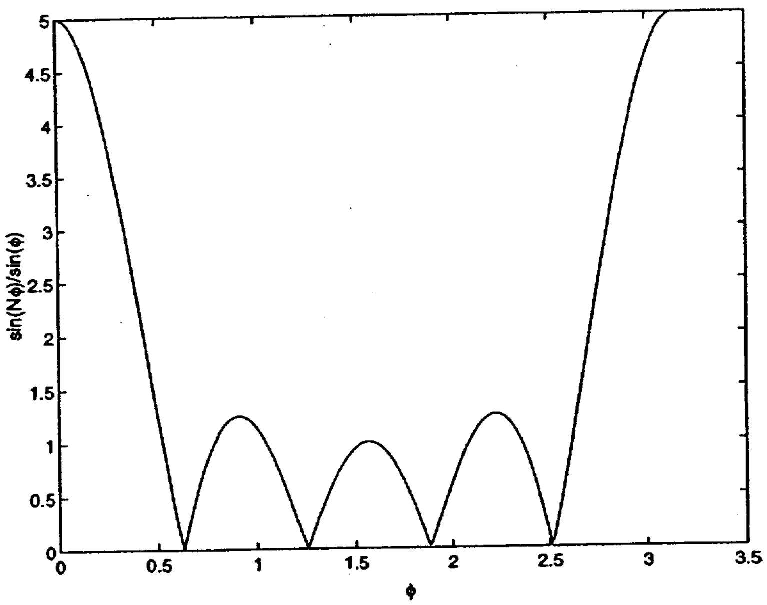


Fig. 3. $|\sin N\phi/\sin\phi|$ versus ϕ of the reverse gain.

Hence, the output voltage, V_{out} is

$$\begin{aligned} V_{\text{out}} &= I_{\text{out}} Z_{\pi d} \\ &= -\tfrac{1}{2} g_m N I_S Z_{\pi g} Z_{\pi d} e^{-jN\phi}. \end{aligned} \tag{9}$$

From (9), the forward transimpedance can be found to be

$$\begin{aligned} Z_f &= \frac{V_{\text{out}}}{I_S} \\ &= -\frac{N}{2} g_m Z_{\pi g} Z_{\pi d} e^{-jN\phi}. \end{aligned} \tag{10}$$

Note that $e^{-jN\phi}$ is the frequency dependent component of Z_f where ϕ is a function of frequency [27] and is valid at frequencies below the cutoff frequency of the line [27]. In a real situation, however, there is a mismatch in impedance between the lines and the terminations at high frequencies (approaching cutoff). This is because the line impedances are frequency-dependent and the terminations are purely resistive.

2) Reverse Transimpedance Gain: The reverse transimpedance gain of the DA is then defined as the ratio between the voltage that appears at the drain termination and the input current. Under a similar analysis as the forward gain where superposition was used, but this time with the output at the drain termination, the drain current is found to be

$$I_{\pi d} = -\frac{1}{2} g_m V_{\text{in}} \frac{\sin \frac{N}{2}(\phi_g + \phi_d)}{\sin \frac{1}{2}(\phi_g + \phi_d)} \exp\left[-j\frac{N}{2}(\phi_g + \phi_d)\right]. \tag{11}$$

With a similar imposed condition of phase matching $\phi_g = \phi_d = \phi$, (11) becomes,

$$I_{\pi d} = -\frac{1}{2} g_m V_{\text{in}} \frac{\sin N\phi}{\sin \phi} e^{-jN\phi}. \tag{12}$$

Thus, reverse transimpedance gain is

$$Z_r = \frac{V_{\pi d}}{I_S} = \frac{Z_{\pi d} I_{\pi d}}{V_{\text{in}}/Z_{\pi d}} = -\frac{1}{2} g_m Z_{\pi g} Z_{\pi d} \frac{\sin N\phi}{\sin \phi} e^{-jN\phi}. \tag{13}$$

Fig. 3 shows the $|\sin N\phi/\sin \phi|$ as a function of ϕ. It is worthwhile to note that the DA has a response similar to a directional coupler with the load placed at the coupled port and the drain termination placed at the isolation port.

B. Noise Analysis

The method of noise analysis is similar to that done by Aitchison [28]; but, it differs in that here the equivalent input noise current density was derived with a current source as input, as opposed to Aitchison's noise figure for a 50 Ω matched system. The MESFET noise sources used in this paper are correlated in contrast to the uncorrelated noise sources used by Aitchison.

The desired result of the noise analysis in this section is to derive an expression for the equivalent current noise density. This is done by breaking the noise contribution into many independent noise power densities, then summing the contribution to get the total noise power density. For each of the current noise sources, there is a corresponding output noise voltage. The equivalent current noise density at the input is derived by dividing the output voltage noise density by the transimpedance gain of the preamplifier. Then the total equivalent input noise current density is obtained by adding the contribution of the individual noise densities. There are several noise sources in the DA, and they are listed as follows:

- noise from the gate termination, $Z_{\pi g}$;
- noise from the drain termination, $Z_{\pi d}$;
- noise from the N FET's.

The terminations are resistive and therefore the single-sided thermal noise current densities are given by

$$|i_{Z\pi g}|^2 = \frac{4kT_0}{Z_{\pi g}}$$
$$|i_{Z\pi d}|^2 = \frac{4kT_0}{Z_{\pi d}} \tag{14}$$

where k = Boltzmann's constant = 1.38×10^{-23} J/°K, and T_0 = Absolute temperature = 290°K.

The Van Der Ziel model [29], [30] for the noise sources of a transistor was used and is shown in Fig. 2(a). The single-sided noise current densities are given by

$$\begin{aligned}|i_g|^2 &= \frac{4kT_0\omega^2 C_{gs}^2 R}{g_m}\\ |i_d|^2 &= 4kT_0 g_m P\\ i_d^* i_g &= jC\sqrt{|i_d|^2|i_g|^2}\\ &= j4kT_0\omega C_{gs} C\sqrt{RP}\end{aligned} \tag{15}$$

where R, P, and C are noise parameters of the transistor [30]. This noise model was chosen for its simplicity and ease in the noise analysis using the simplified transistor model shown in Fig. 2(a).

1) Noise From Gate Termination $Z_{\pi g}$: From Fig. 2(b), it can be seen that half of the noise current, $i_{Z\pi g}$, is absorbed by $Z_{\pi g}$ itself, while the other half travels toward the input. The output voltage produced by this current wave is formed by a reverse transimpedance gain, Z_r. This current wave continues to travel toward the input. It then gets totally reflected since the input is effectively an open circuit (current source). The output voltage produced by this reflected wave is formed by a forward transimpedance gain, Z_f. The total voltage appearing at the output due to a source is

$$v_{o1} = Z_f \frac{i_{Z\pi g}}{2} e^{-jN\phi} + Z_r \frac{i_{Z\pi g}}{2}. \tag{16}$$

The equivalent current source at the input to produce the same output in (16) is

$$i_{n1} = \frac{v_{o1}}{Z_f} = \frac{i_{Z\pi g}}{2} e^{-jN\phi}\left(1 + \frac{Z_r}{Z_f} e^{jN\phi}\right). \tag{17}$$

One may therefore write the current transfer function $H(f)$ as

$$H(f) = \frac{i_{n1}}{i_{Z\pi g}} = \frac{e^{-jN\phi}}{2}\left(1 + \frac{Z_r}{Z_f} e^{jN\phi}\right). \tag{18}$$

Thus the equivalent input current density is

$$|i_{n1}|^2 = |H(f)|^2 |i_{z\pi g}|^2 = \frac{|i_{z\pi g}|^2}{4}\left|1 + \frac{Z_r}{Z_f} e^{jN\phi}\right|^2. \tag{19}$$

In the midfrequency range of the DA, $Z_r \gg Z_f$ and thus the equivalent input noise current density is $\frac{1}{2}\sqrt{|i_{Z_{\pi g}}|^2}$. In the case of the conventional lumped circuit Fig. 1(a) and (b), all the thermal noise due to R_{bias} and R_f is seen directly at the input that is $\sqrt{|i_{R_{\text{bias}}}}$ and $\sqrt{|i_{R_f}|^2}$, respectively.

2) Noise From Drain Termination $Z_{\pi d}$: Half of the drain termination noise current gets absorbed by the termination itself while the other half travels down the drain line and gets dissipated in the output producing a voltage noise source, v_{o2}

$$v_{o2} = \frac{i_{Z_{\pi d}}}{2} e^{-jN\phi} Z_{\pi d}. \tag{20}$$

Hence the equivalent input current for this output voltage is

$$i_{n2} = \frac{v_{o2}}{Z_f} = \frac{Z_{\pi d}}{Z_f}\frac{i_{Z_{\pi d}}}{2} e^{-jN\phi}. \tag{21}$$

The equivalent input noise power density is

$$|i_{n2}|^2 = \left|\frac{Z_{\pi d}}{Z_f}\right|^2 \frac{|i_{Z_{\pi d}}|^2}{4}. \tag{22}$$

3) Noise from the MESFET Transistors: The total equivalent input noise density of N FET's equals the square sum of each of the noise densities referred to the input. Consider the noise sources of rth stage FET. There are two noise sources in a FET, gate noise source and drain noise source.

The gate noise current wave can be broken into three components. The first component is the forward traveling wave which will produce an output voltage with a forward transimpedance gain. The second is the reverse traveling wave which will produce an output voltage with a reverse transimpedance gain. Finally, the third component is the reflected wave. It is due to the reverse wave reflecting off the open circuit at the input. This wave will also produce an output voltage with a forward transimpedance gain.

Defining the output current due to the forward traveling wave as $i_{\text{out}}^{\text{for}}$ results in the following expression:

$$\begin{aligned}i_{\text{out}}^{\text{for}} &= \tfrac{1}{2}(i_r^{\text{for}} \exp[-j(N - r - \tfrac{1}{2})\phi_d]\\ &\quad + i_{r+1}^{\text{for}} \exp[-j(N - r - \tfrac{3}{2})\phi_d]\\ &\quad + \cdots + i_N^{\text{for}} \exp(-j - \tfrac{1}{2}\phi_d))\end{aligned} \tag{23}$$

where

$$\begin{aligned}i_{r+i}^{\text{for}} &= \text{drain current due to forward traveling}\\ &\qquad \text{wave at } (r+i)\text{th cell's gate}\\ &= g_m v_{gs_{r+1}} = \tfrac{1}{2} g_m i_{gr} Z_{\pi g} e^{-j(i)\phi_g},\\ &\qquad i = 0, 1, 2, \cdots, (N - r).\end{aligned} \tag{24}$$

Using a similar method as was used to derive the forward gain

$$i_{\text{out}}^{\text{for}} = \frac{(N - r + 1)}{4} g_m i_{gr} Z_{\pi g} \exp\left[-j\left(N - r + \frac{1}{2}\right)\phi\right]. \tag{25}$$

Similarly, the output current due to the reverse traveling wave, $i_{\text{out}}^{\text{rev}}$, is

$$\begin{aligned}i_{\text{out}}^{\text{rev}} &= \frac{1}{4} g_m i_{gr} Z_{\pi g} \frac{\sin(r-1)\phi}{\sin\phi}\\ &\quad \cdot \exp\left[-j\left(N + r - \frac{1}{2}\right)\phi\right] e^{j\phi(r-1)}.\end{aligned} \tag{26}$$

The output current due to the reflected gate noise current wave can be shown to be

$$i_{\text{out}}^{\text{ref}} = \frac{1}{4} g_m i_{gr} Z_{\pi g} \exp\left[-j\left(N + r - \frac{1}{2}\right)\phi\right] N. \tag{27}$$

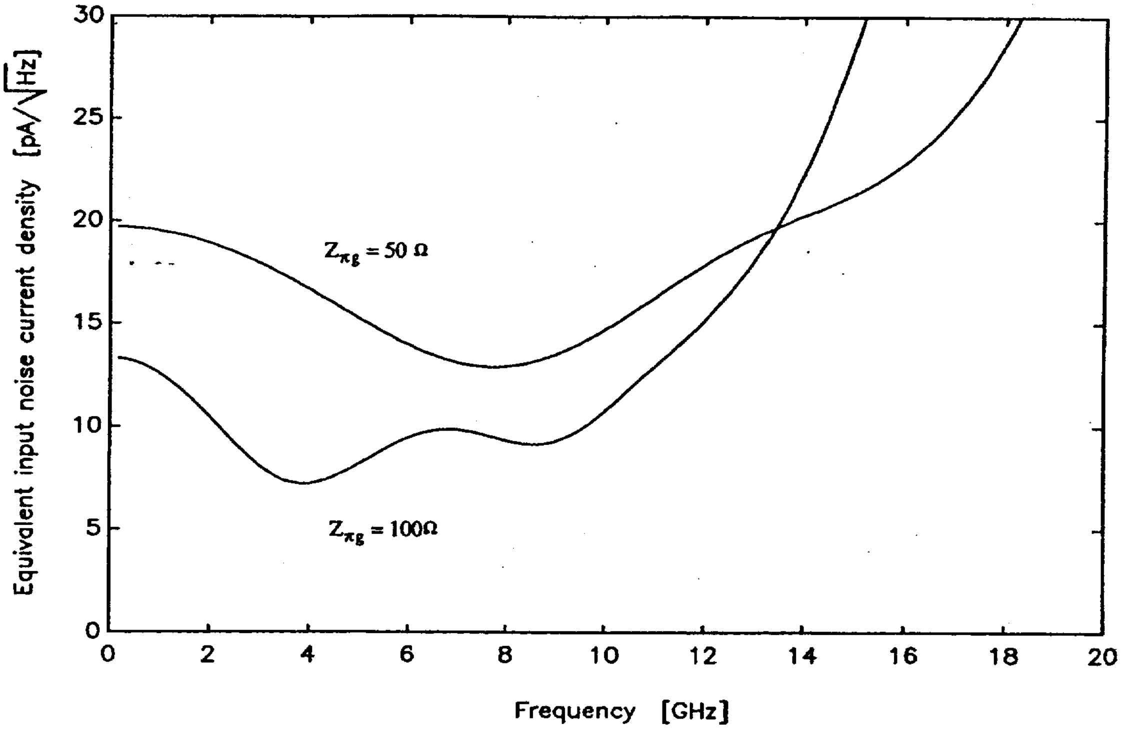


Fig. 4. Total equivalent noise current density for two different values of $Z_{\pi g}$.

So the total output current due to the gate noise of rth is

$$\begin{aligned} i_{\text{out}}^{gr} &= i_{\text{out}}^{\text{for}} + i_{\text{out}}^{\text{ref}} + i_{\text{out}}^{\text{rev}} \\ &= \frac{1}{4} g_m i_{gr} Z_{\pi g} \exp\left[-j\left(N + r - \frac{1}{2}\right)\phi\right] \\ &\quad \cdot \left(N + (N - r + 1)e^{j(2r-1)\phi} \right. \\ &\quad \left. + \frac{\sin(r-1)\phi}{\sin\,\phi} e^{(r-1)\phi}\right). \end{aligned} \tag{28}$$

The resulting output voltage is

$$\begin{aligned} v_{\text{out}}^{gr} &= i_{\text{out}}^{gr} Z_{\pi d} \\ &= \frac{1}{4} g_m i_{gr} Z_{\pi g} Z_{\pi d} N \exp\left[-j\left(N + r\frac{1}{2}\right)\phi\right] \\ &\quad \cdot \left(1 + \frac{(N - r + 1)}{N} e^{j(2r-1)\phi} \right. \\ &\quad \left. + \frac{1}{N} \frac{\sin(r-1)\phi}{\sin\,\phi} e^{j(r-1)\phi}\right). \end{aligned} \tag{29}$$

Substituting the expression for transimpedance gain in (10) into (29), and writing the right-hand bracketed in rectangular form

$$v_{\text{out}}^{gr} = \tfrac{1}{2} i_{gr} Z_f \exp[-j(r - \tfrac{1}{2})\phi](A(r,\phi) + jB(r,\phi)) \tag{30}$$

where

$$\begin{aligned} A(r,\phi) &= 1 + \frac{(N - r + 1)}{N} \cos(2r-1)\phi \\ &\quad + \frac{1}{N} \frac{\sin(r-1)\phi}{\sin\,\phi} \cos(r-1)\phi \end{aligned}$$

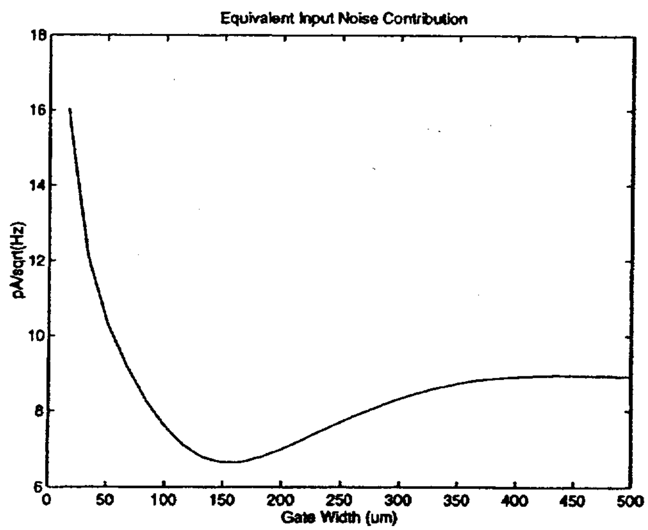


Fig. 5. Equivalent input current noise density versus gate width.

$$\begin{aligned} B(r,\phi) &= \frac{(N - r + 1)}{N} \sin(2r-1)\phi \\ &\quad + \frac{1}{N} \frac{\sin(r-1)\phi}{\sin\,\phi} \sin(r-1)\phi. \end{aligned} \tag{31}$$

For the drain noise, the analysis is much simpler. Half of the drain noise current will travel toward the termination and be absorbed, while the other half will travel toward the load and produce a corresponding output voltage. So assuming again

that phase matching condition still holds, the total current due to the drain noise of the rth stage can be written as

$$i_{\text{out}}^{dr} = \tfrac{1}{2} i_{dr} \exp[-j(N - r + \tfrac{1}{2})\beta]. \tag{32}$$

Hence the corresponding output voltage is

$$\begin{aligned} v_{\text{out}}^{dr} &= v_{\text{out}}^{dr} Z_{\pi d} \\ &= \tfrac{1}{2} i_{dr} Z_{\pi d} \exp[-j(N - r + \tfrac{1}{2})\phi]. \end{aligned} \tag{33}$$

Thus the total output noise voltage due to the FET noise of the rth stage is as follows:

$$\begin{aligned} v_{\text{out}}^{r} &= v_{\text{out}}^{gr} + v_{\text{out}}^{dr} \\ &= \tfrac{1}{2} i_{gr} Z_f \exp[-j(r - \tfrac{1}{2})\phi](A(r,\phi) + jB(r,\phi)) \\ &\quad + \tfrac{1}{2} i_{dr} Z_{\pi d} \exp[-j(N - r + \tfrac{1}{2})\phi]. \end{aligned} \tag{34}$$

The equivalent input noise current due to the output noise voltage in (34) is

$$\begin{aligned} |i_n^r|^2 &= (A(r,\phi)^2 + B(r,\phi)^2)\frac{|i_{gr}|^2}{4} \\ &\quad + \left|\frac{Z_{\pi d}}{Z_f}\right|^2 \frac{|i_{dr}|^2}{4} + \frac{1}{2} \\ &\quad \cdot \operatorname{Re}\left(i_{gr} i_{dr}^* \frac{Z_{\pi d}^*}{Z_f^*} (A(r,\phi) + jB(r,\phi)) e^{j(N-2r+1)\phi} \right). \end{aligned} \tag{35}$$

So assuming that all gate and drain current sources are the same for all stages, i.e., $i_{gr} = i_g$, and $i_{dr} = i_d$, the total equivalent input noise power density of all FET's in the N stages is

$$\begin{aligned} |i_{n3}|^2 &= \sum_{r=1}^{N} |i_n^r|^2 \\ &= \frac{|i_g|^2}{4} \sum_{r=1}^{N} (A(r,\phi)^2 + B(r,\phi)^2) \\ &\quad + \frac{|i_d|^2}{4} \left|\frac{Z_{\pi d}}{Z_f}\right|^2 N + \frac{1}{2} \sum_{r+1}^{N} \\ &\quad \cdot \operatorname{Re}\left(i_g i_d^* \frac{Z_{\pi d}^*}{Z_f^*} (A(r,\phi) + jB(r,\phi)) e^{j(N-2r+1)\phi} \right). \end{aligned} \tag{36}$$

Equations (19), (22), and (36) give the total equivalent input noise current density, $\sqrt{|i_n|^2}$, shown in (37), at the bottom of the page. Equation (37) can be written completely in terms of ω by assuming that $\phi = \beta d = k = \omega\sqrt{L_g C_{gs}}$, where L_g and C_{gs} are the gate line inductance and capacitance of the DA.

The plots of total equivalent input noise density versus frequency are shown in Fig. 4 for two impedances $Z_{\pi g} = 50\ \Omega$ and $100\ \Omega$. The following parameters were used in the calculation: $Z_{\pi d} = 50\ \Omega, L_g = 0.5$ nH $(Z_{\pi g} = 50\ \Omega), L_g = 2$ nH $(Z_{\pi g} = 100\ \Omega), n = 5, C_{gs} = 200$ fF, $g_m = 20$ mS. When $Z_{\pi g} = 100\ \Omega$, the equivalent input noise current density is as much as 5 pA/$\sqrt{\text{Hz}}$ less than that of the case when $Z_{\pi g} = 50\ \Omega$. $Z_{\pi g} = \sqrt{L_g/C_{gs}}$ should be made as large as possible to reduce the low frequency noise contribution. Increasing the gate line inductance increases the impedance of the gate line and decreases the cutoff frequency, $f_c = 1/\pi\sqrt{L_g C_{gs}}$. Thus there is a tradeoff between noise performance and bandwidth of the DA preamplifier.

In (37), a plot of total equivalent input current noise density versus gate width can be obtained, when expressing $C_{gs} = C_{gso}W$ and $g_m = g_{mo}W$. C_{gso} (1 nF/m) is the capacitance per unit gate width and g_{mo} (125 S/m) is the transconductance per unit gate width and the gate width of the transistor is W. From Fig. 5, it can be seen that the gate width of 150 μm produces the lowest noise.

C. Front-End Tuning

As the data rate increases, receivers with high sensitivity must have extended bandwidths of operation into the gigahertz region. A tuning network placed after the photodiode can significantly help reduce the noise at high frequencies [1], [2]. One of the approaches is to use the optimal admittance seen by the first stage transistor to obtain the best noise figure [1]. This is done by designing a tuning network (including the photodiode) whose output admittance is equal to the optimal input admittance for the transistor in order to get the minimum noise figure, which is given by the manufacturer's data sheet.

In a DA design, however, it is difficult to apply the same method because of the distributed nature of the structure. It is for this reason that the design of the front-end tuning network in this paper is approached from the theory of network analysis. Fig. 6(a) shows a noiseless amplifier with its equivalent input noise power density, $|i_{\text{amp}}|^2$.

Fig. 6(b) shows a general low-pass filter structure with its $[ABCD]$ matrix representation. The filter is driven by current source with the photodiode capacitance absorbed by the filter shown in Fig. 6(c). The load of the filter is the input impedance of the amplifier Z_{amp}, which is real in a DA design. For the $[abcd]$ matrix of the lossless filter tuning network in Fig. 6(b)

$$\begin{aligned} \sqrt{|i_n|^2} &= \sqrt{|i_{n1}|^2 + |i_{n2}|^2 + |i_{n3}|^2} \\ &= \sqrt{\begin{aligned} &\frac{|i_{Z\pi g}|^2}{4}\left|1 + \frac{Z_r}{Z_f} e^{jN\phi}\right|^2 + \frac{1}{|Z_f|^2}\frac{|i_{Z_{\pi d}}|^2}{4} Z_{\pi d}^2 + \frac{|i_g|^2}{4} \sum_{r+1}^{N} (A(r,\phi)^2 + B(r,\phi)^2) \\ &+ \frac{|i_d|^2}{4}\left|\frac{Z_{\pi d}}{Z_f}\right|^2 N + \frac{1}{2}\sum_{r+1}^{N} \operatorname{Re}\left(i_g i_d^* \frac{Z_{\pi d}^*}{Z_f^*}(A(r,\phi) + jB(r,\phi)) e^{j(N-2r+1)\phi}\right) \end{aligned}} \end{aligned} \tag{37}$$

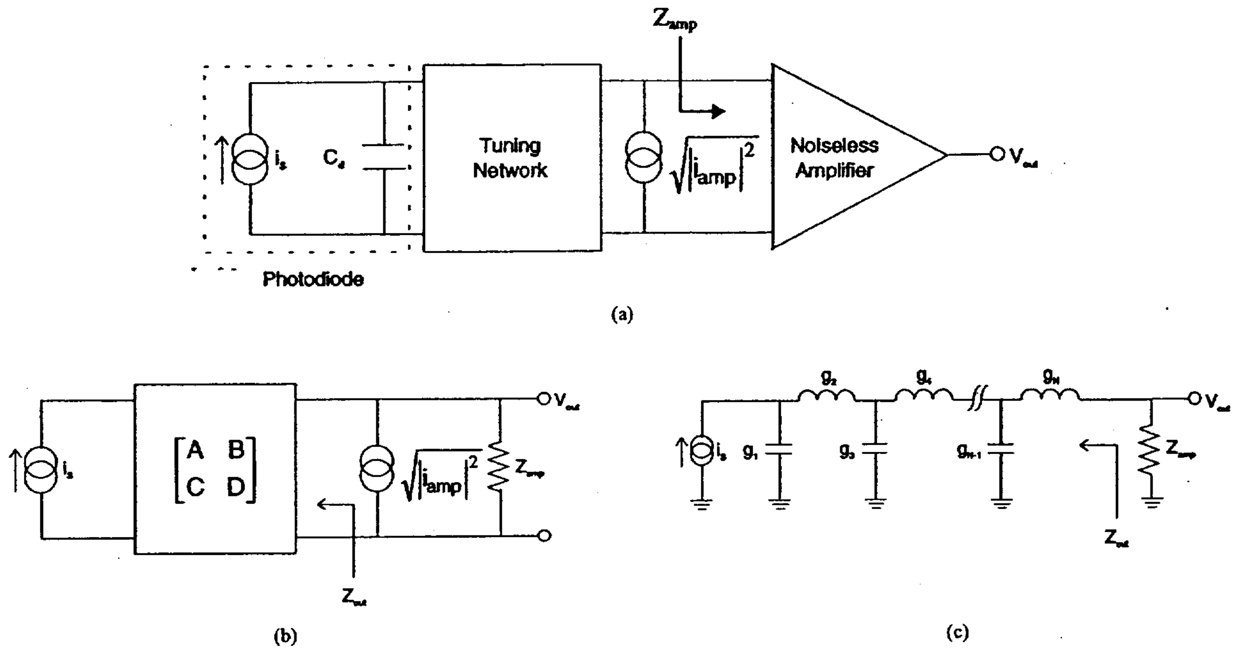


Fig. 6. (a) Noiseless amplifier and tuning network, (b) general filter structure with $[ABCD]$ matrix representation, and (c) general lossless low-pass filter with a single load and infinite source impedance.

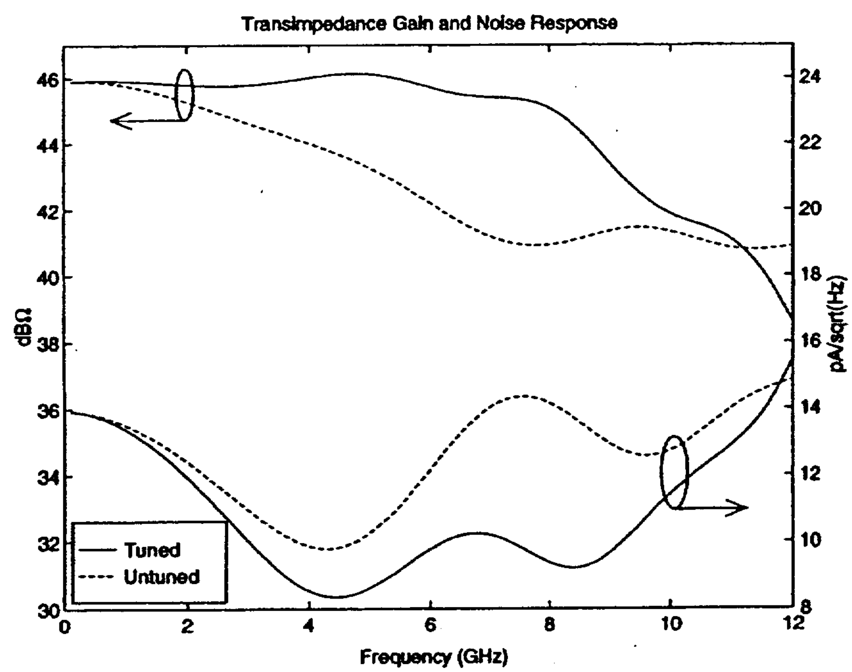


Fig. 7. Theoretical tuned and untuned transimpedance gain and noise response.

and the load

$$\begin{bmatrix} a & b \\ c & d \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 1/Z_{\text{amp}} & 1 \end{bmatrix} = \begin{bmatrix} A + B/Z_{\text{amp}} & B \\ C + D/Z_{\text{amp}} & D \end{bmatrix}. \tag{38}$$

The overall transimpedance gain of the loaded structure is then

$$Z_t = \frac{1}{c} = \frac{1}{C + \dfrac{D}{Z_{\text{amp}}}} \tag{39}$$

and the output impedance of the general filter is

$$Z_{\text{out}} = \frac{D}{C}. \tag{40}$$

When Z_{out} is zero, the noise current source of the amplifier will be shunted by the tuning network to ground. Equation (40) implies that the only way to make Z_{out} zero within the pass band is to make "D" zero or to have "C" approach infinity. The latter condition means that if "C" approaches infinity, Z_t will approach zero [see (39)]. This condition will not occur

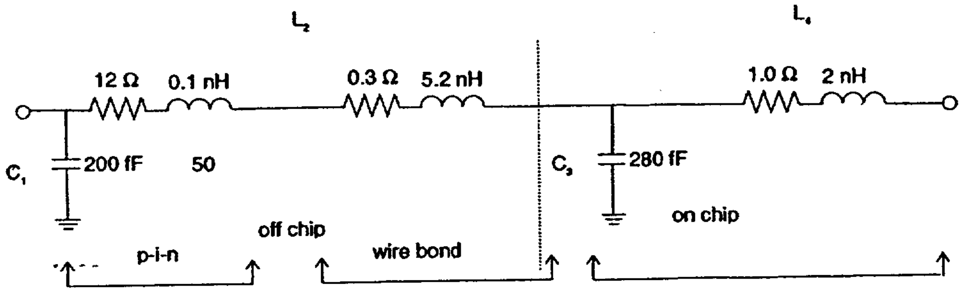


Fig. 8. Actual tuning network and component values used in the design.

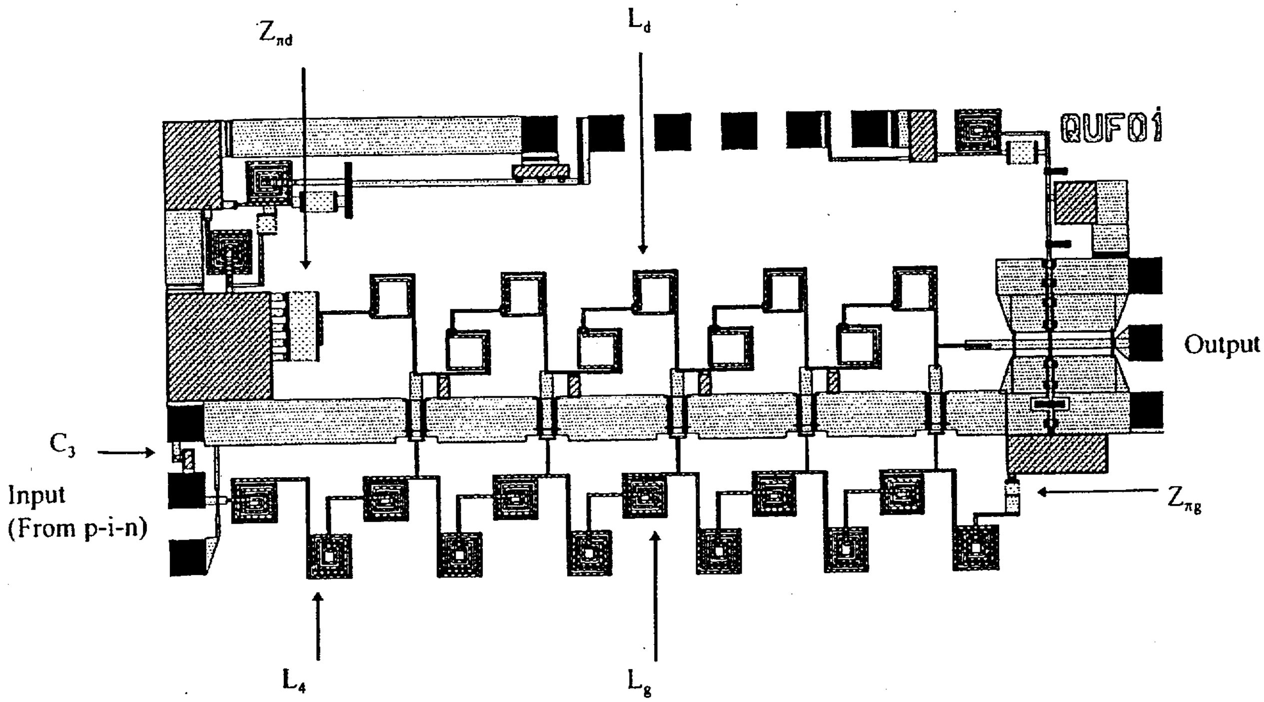


Fig. 9. Distributed preamplifier layout. Note the large inductance of L_g is to obtain $Z_{\pi g} = 100\ \Omega$ and the on-chip front-end tuning circuit C_3 and L_4.

TABLE I
INITIAL ESTIMATES OF THE FRONT-END TUNING NETWORK

Reference[31] 1 dB equi-ripple filter Low pass proto-type	Values Before Optimization Cutoff frequency 8 GHz Z_{amp}=100 Ω	Values after Optimization Cutoff frequency 8 GHz Z_{amp}=100 Ω
g_1=1.2817	C_1= 255 fF	C_1= 200 fF
g_2=1.9093	L_2= 3.8 nH	L_2= 5.3 nH
g_3=1.4126	C_3= 281 fF	C_3= 280 fF
g_4=1.0495	L_4= 2.1nH	L_4= 2 nH

since the transimpedance of the overall structure is designed to have a low-pass frequency response.

Fig. 6(c) shows a typical general lossless low-pass filter of Nth order. The $[ABCD]$ matrices of a capacitor and an inductor are defined as

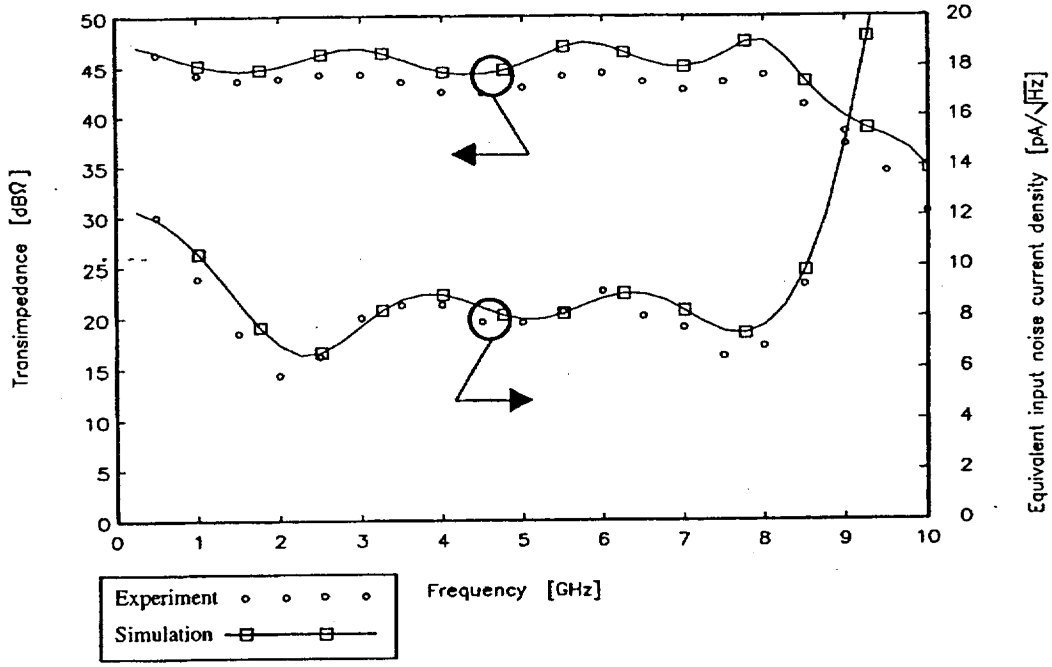
$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{Capacitor}} = \begin{bmatrix} 1 & 0 \\ sg_i & 1 \end{bmatrix} \tag{41}$$

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{Inductor}} = \begin{bmatrix} 1 & sg_j \\ 0 & 1 \end{bmatrix} \tag{42}$$

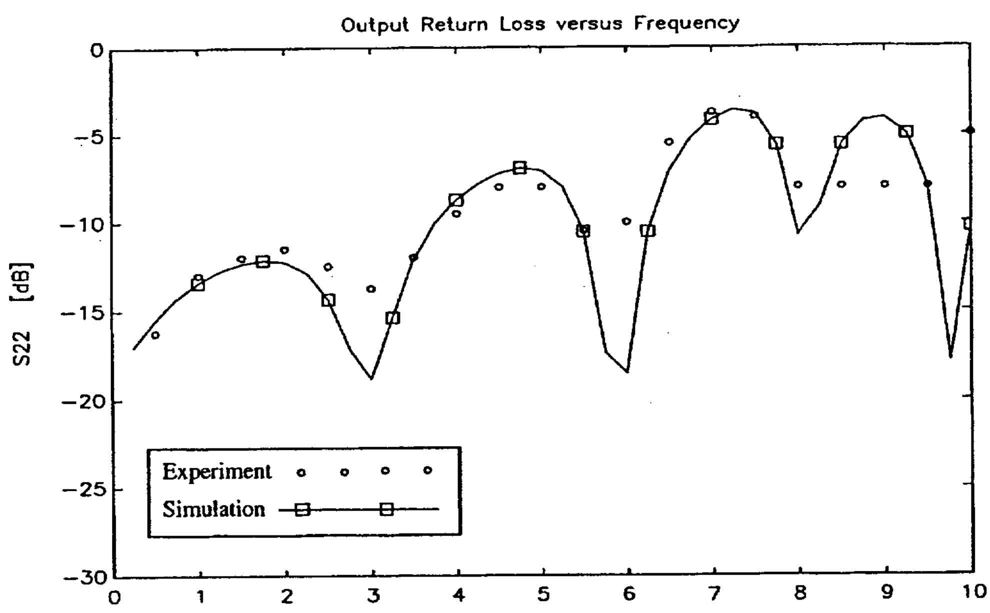
where $i, j = 1, 2 \cdots N, i \neq j$. The overall $[ABCD]$ matrix is found by multiplying N matrices together using (41) and (42). The following polynomials can be obtained for the "C" and "D" elements of the $[ABCD]$ matrix

$$C = \left(\sum_{i=1}^{N/2} k_{2i-1}(s)^{2i-1} \right) = k_1 s + k_3 s^3 + k_5 s^5 \cdots s^{N-1} k_{N-1} \tag{43}$$

$$D = 1 + \left(\sum_{i=1}^{N/2} k_{2i}(s)^{2i} \right) = 1 + k_2 s^2 + k_4 s^4 + k_6 s^6 + \cdots + s^N k_N \tag{44}$$



(a)



(b)

Fig. 10. (a) The simulated and measured transimpedance and equivalent input noise current densities of a distributed preamplifier. (b) Output return loss of the DA preamplifier.

where $k_m, m = 1, 2 \cdots N$, is the summations of products and cross products of g_j and g_i.

It can be seen that Z_{out} has N zeros and $N-1$ poles since there are N frequencies at which "D" is zero and $N-1$ frequencies at which "C" is zero. Note that when "C" becomes zero, Z_{out} will approach infinity. This simply implies that the tuning network has no effect on the noise performance at these frequencies. More importantly, this means that the noise performance will never get worse with the tuning network. At worst, it gives the same result as the case with no tuning network. However, when "D" is zero, Z_{out} is zero and the noise $\sqrt{|i_{\text{amp}}|^2}$ is minimized.

In a practical design, the capacitance of the photodiode is absorbed into the filter by replacing g_1 with the capacitance of

the photodiode. The whole structure is then optimized with the p-i-n resistance included to equalize the rolloff characteristics of the preamplifier. Fig. 7 shows the effect of front-end tuning on the theoretical transimpedance and noise response of the same DA.

The tuning network used in this design is a singly terminated fourth-order Tchebyshev filter [31]. The first estimate of the component values was achieved by the standard filter design technique of single-ended termination of 1 dB ripple, normalized to impedance of $Z_{\text{amp}} = 100\ \Omega$ and frequency of 8 GHz. The following table summarizes the initial estimates of the component values of the tuning network.

Fig. 8 shows the actual tuning network and the component values used in the design. The initial estimated values of the components, as summarized in Table I, and the model of the p-i-n are then inserted in Touchstone by HP/Eesof for optimization in terms of gain flatness and noise performance. The optimal values obtained with Touchstone by HP/Eesof optimization routine are also included. The length of the wire bond required to meet the designed value was calculated to be 4.5 mm for a 0.7 mil diameter bond wire.

III. Results

A monolithic distributed preamplifier was designed, Fig. 9, using Touchstone by HP/Eesof with $Z_{\pi g} = 100\ \Omega, Z_{\pi d} = 50\ \Omega$ and was constructed using a 0.8 μm self-aligned gate process at NT. The NT MESFET had the following process parameters: $V_p = -1.2$ V, $I_{dss} = 120$ A/mm, $f_t = 20$ GHz and $F_{\min} < 1$ dB up to 5 GHz. The optimal gate width of the MESFET for the best noise performance was obtained to be $w_{\text{opt}} = 150\ \mu$m and was used in the design. With $w_{\text{opt}} = 150\ \mu$m, the cutoff frequency for the distributed preamplifier was $f_c = 16$ GHz. L_g of the gate line had a measured self resonance of greater than 20 GHz. The preamplifier chip was then wire bonded to a 35 μm NT p-i-n. The bond wire inductance was $L_s = 5.3$ nH, which was used as inductive tuning. Measured results of the distributed preamplifier are shown in Fig. 10. It can be seen that the receiver has an equivalent input current density of 8 pA/$\sqrt{\text{Hz}}$, a transimpedance gain of 46 dBΩ, and a 3 dB bandwidth of 8 GHz. Using 8 pA/$\sqrt{\text{Hz}}$ and 8 GHz bandwidth, the predicted receiver sensitivity can be shown to be -21 dBm [25] for a 10 Gb/s system. This is the best sensitivity that one could expect for this preamplifier since there would be a penalty associated due to passband ripple. The ripple in the transimpedance gain was measured to be 2 dB and, based on back-simulation, the penalty due to eye closure is no more that 5 dB. The passband ripple was 1 dB higher than designed and was a result of the sheet resistance for the resistors being 20% higher. Thus, the gate line was no longer matched and a standing wave existed on it. The higher sheet resistance also resulted in a poorer return loss at the output since $Z_{\pi d}$ was larger. The measured return loss was 7 dB at 5 GHz, as seen in Fig. 10(b). These results could be improved upon by a second iteration of the preamplifier chip.

IV. Conclusion

A low noise broadband GaAs MESFET monolithic distributed preamplifier has been presented. The theory of noise in a distributed MESFET preamplifier is developed. The high gate line matching impedance improves the noise performance of the distributed preamplifier. With the increase in gate line impedance comes a reduction in the bandwidth of the preamplifier. The preamplifier was optimized for the best total equivalent input noise current density by adjusting the width to $w_{\text{opt}} = 150\ \mu$m. Tchebyshev front-end circuit tuning was used to reduce the high frequency noise. This filter-type tuning network included on-chip components, wire bond inductance, and the parasitics of the photodiode. A monolithic distributed preamplifier demonstrated -3 dB bandwidth of greater than 8 GHz and an average input noise current density of 8 pA/$\sqrt{\text{Hz}}$. The predicted best receiver sensitivity at 10 Gb/s is -21 dBm. To our knowledge, the preamplifier described has the lowest noise performance in this band based on 0.8 μm GaAs MESFET technology.

Acknowledgment

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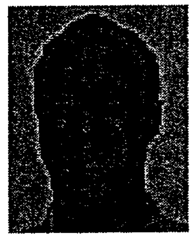
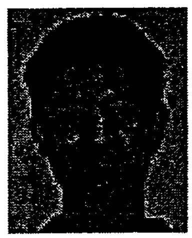


Al P. Freundorfer (M'90) received the B.A.Sc., M.A.Sc., and Ph.D. degrees all from the University of Toronto, ON, Canada in 1981, 1983, and 1989, respectively.

In 1990 he joined the Department of Electrical Engineering at Queen's University, Kingston, ON Canada. Since then he has done work in nonlinear optics of organic crystals, coherent optical network analysis as well as microwave integrated circuits. Currently he is focusing his attention on monolithic microwave circuits used in lightwave systems with bit rates in excess of 20 Gb/s and on monolithic millimeter wave integrated circuits used in wireless communications.



Thé Linh Nguyen received the B.A.Sc. and M.A.Sc. degrees in 1993 and 1995 at the University of Waterloo and Queen's University, respectively. For his Master's thesis, he designed a monolithic balanced distributed preamplifier for high-speed coherent optical communications.

He is currently a member of scientific staff at Bell-Northern Research Ltd. located in Ottawa, ON Canada. His work involves design of transmitter and receiver, and MMIC design for high-speed optical systems.

TAB 11

United States Patent [19]

Ash

[11] 4,162,452

[45] Jul. 24, 1979

[54] CHANNEL SELECTION FOR A TELEVISION RECEIVER HAVING LOW-GAIN HIGH FREQUENCY RF-IF SECTION

[75] Inventor: Darrell L. Ash, Garland, Tex.

[73] Assignee: Texas Instruments Incorporated, Dallas, Tex.

[21] Appl. No.: 813,137

[22] Filed: Jul. 5, 1977

[51] Int. Cl.² H04B 1/16

[52] U.S. Cl. 325/373; 325/472; 325/464

[58] Field of Search 358/191; 325/371, 373, 325/388, 375, 472, 473, 464

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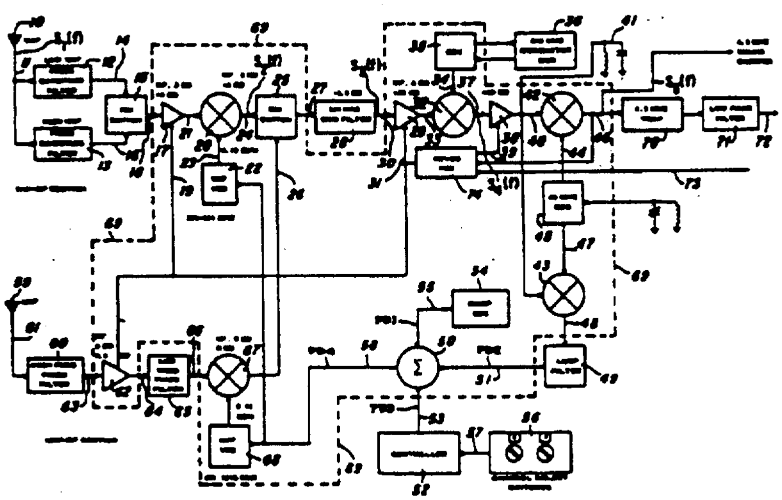
3,622,891 11/1971 Leland 325/472

Primary Examiner—Robert L. Richardson
Attorney, Agent, or Firm—Melvin Sharp; Richard L. Donaldson; William E. Hiller

[57] ABSTRACT

A channel selector for a television receiver having a low gain, high-frequency radio frequency-intermediate frequency section as combined with a relatively high-gain, lower intermediate frequency in the intermediate frequency section to reduce noise, such as intermodulation distortion and cross-modulation distortion. The channel selector comprises a frequency spectrum filter in the RF section, a mixer which preferably includes a metal semiconductor field effect transistor, i.e. MESFET, and a channel selecting filter. The frequency spectrum filter is coupled to the antenna of the television receiver and filters a frequency spectrum of radio frequency signals as received from the antenna, such as a plurality of television channels. The mixer has an input coupled to the frequency spectrum filter and frequency shifts selected channels of the frequency spectrum to a predetermined relatively high intermediate frequency. The RF section has a low gain RF amplifier connected between the frequency spectrum filter and the mixer or may have no RF amplifier at all. The channel selecting filter has an input coupled to the mixer and filters the selected channel at the predetermined intermediate frequency. The total gain from the RF section through the channel selecting filter is no larger than necessary in order to obtain a desired system noise figure. Another mixer is provided in the IF section, this IF mixer being coupled to the output of the channel selecting filter. The IF mixer receives the filtered selected channel at the relatively high intermediate frequency from the channel selecting filter and frequency shifts the filtered selected channel to a substantially lower second intermediate frequency. An IF amplifier having a relatively high gain is connected to the output of the IF mixer and adds gain to the filtered selected channel at the lower second intermediate frequency, the added gain being substantially in excess of the total gain provided before the channel selecting filter.

8 Claims, 40 Drawing Figures



U.S. Patent Jul. 24, 1979 Sheet 1 of 16 4,162,452

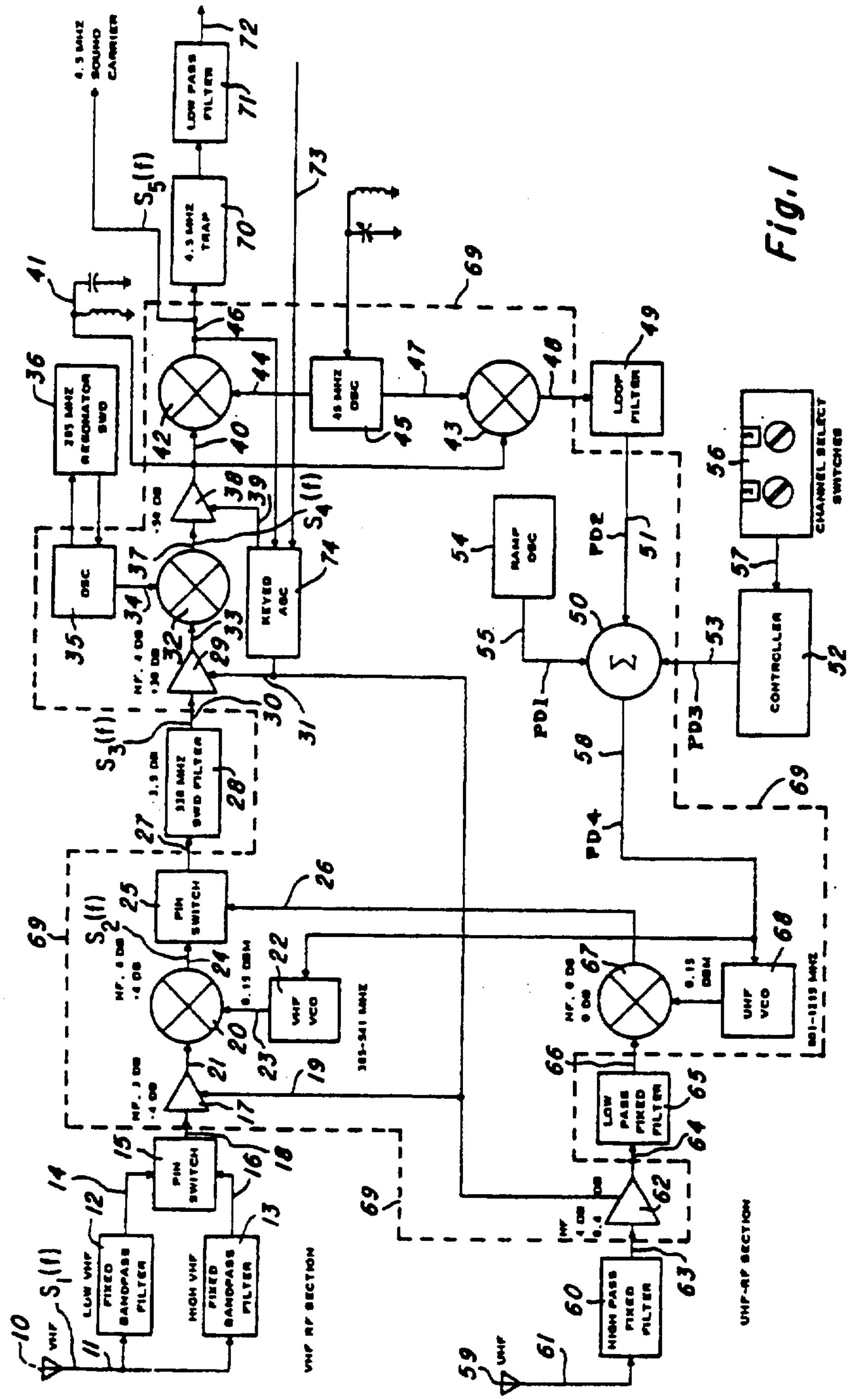
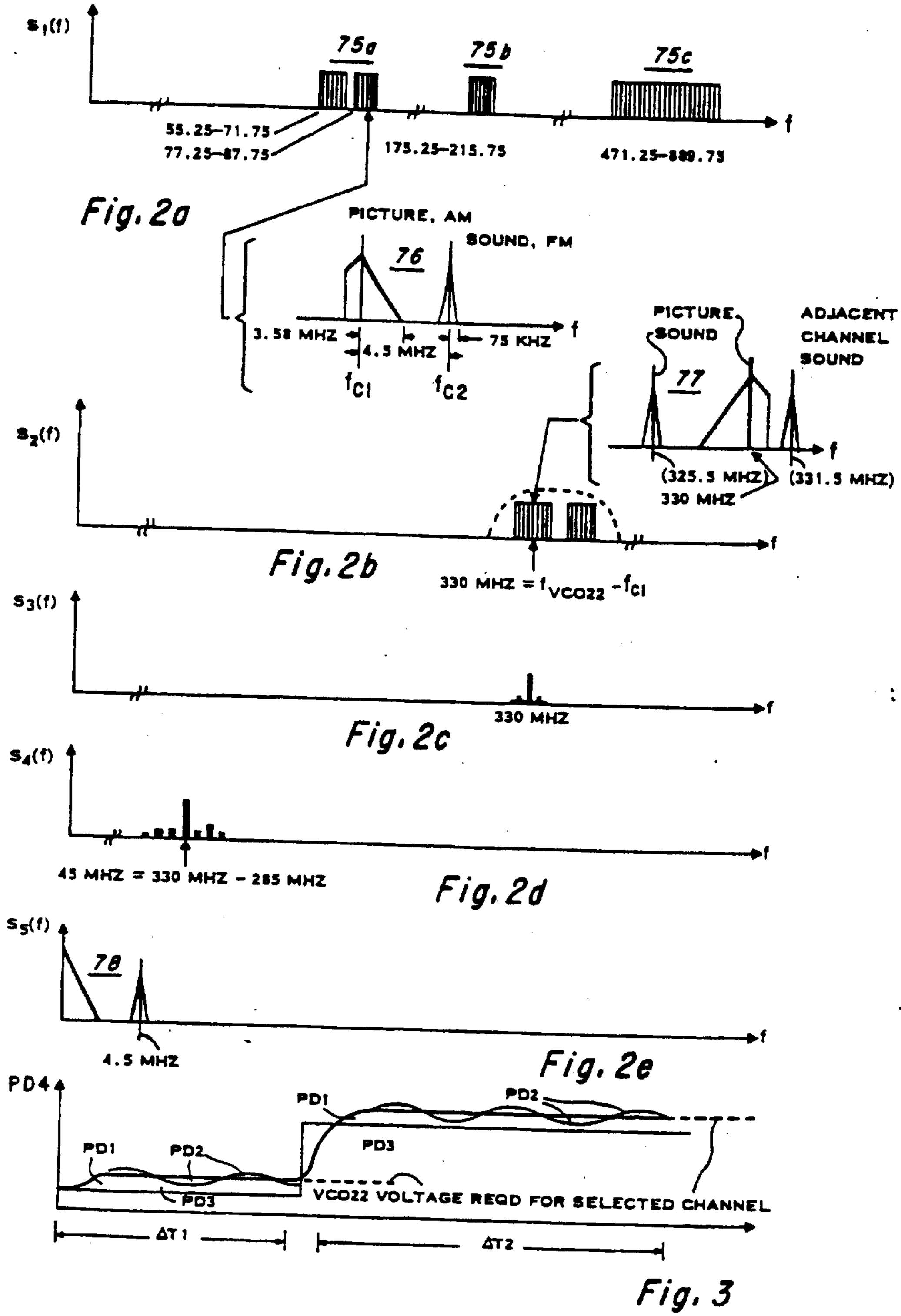


Fig. 1



$S_1(f)$
75a
75b
75c
f
55.25–71.75
77.25–87.75
175.25–215.75
471.25–889.75
Fig. 2a
PICTURE, AM
SOUND, FM
76
3.58 MHZ
4.5 MHZ
75 KHZ
f_{C1}
f_{C2}
PICTURE SOUND
ADJACENT CHANNEL SOUND
77
(325.5 MHZ)
(331.5 MHZ)
330 MHZ
$S_2(f)$
Fig. 2b
330 MHZ = $f_{VCO22} - f_{C1}$
$S_3(f)$
330 MHZ
Fig. 2c
$S_4(f)$
45 MHZ = 330 MHZ − 285 MHZ
Fig. 2d
$S_5(f)$
78
4.5 MHZ
Fig. 2e
PD4
PD1
PD2
PD3
VCO22 VOLTAGE REQD FOR SELECTED CHANNEL
ΔT1
ΔT2

Fig. 3

Fig. 4a

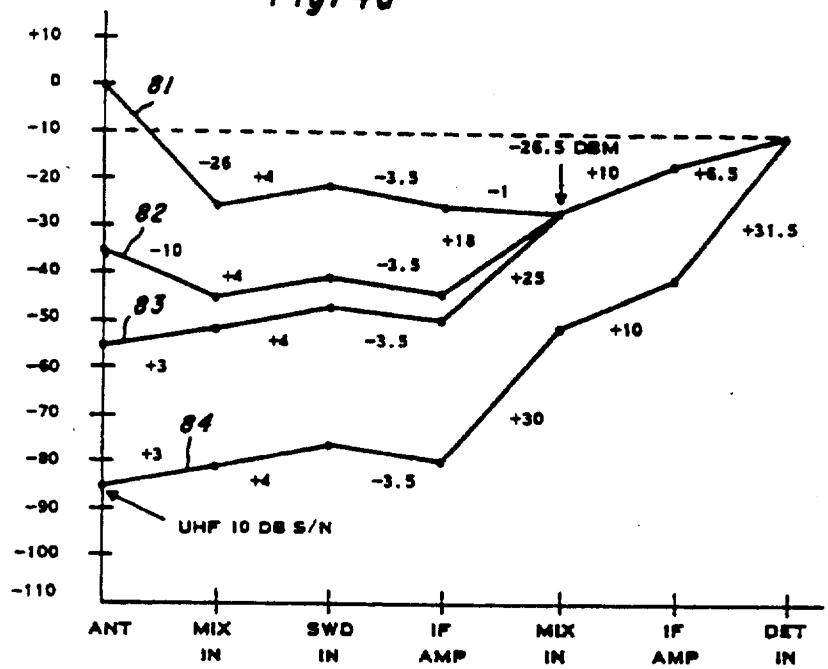
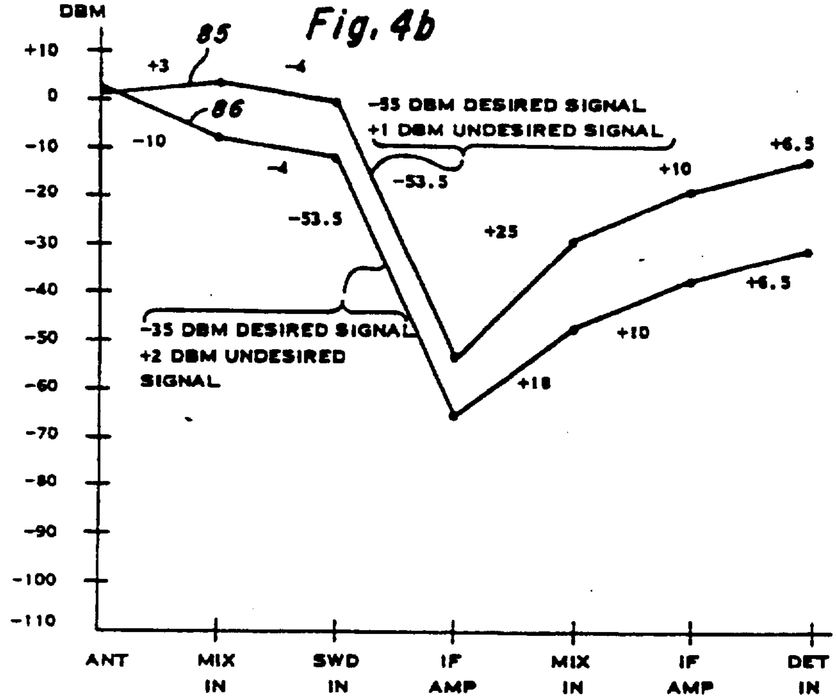

DBM
+10
0
-10
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
81
82
83
84
-26
+4
-3.5
-1
-26.5 DBM
+10
+6.5
+31.5
-10
+18
+25
+3
+30
UHF 10 DB S/N
ANT
MIX IN
SWD IN
IF AMP
MIX IN
IF AMP
DET IN

Fig. 4b


DBM
+10
0
-10
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
85
86
+3
-4
-10
-35 DBM DESIRED SIGNAL
+1 DBM UNDESIRED SIGNAL
-53.5
+25
+10
+6.5
-35 DBM DESIRED SIGNAL
+2 DBM UNDESIRED SIGNAL
+18
ANT
MIX IN
SWD IN
IF AMP
MIX IN
IF AMP
DET IN

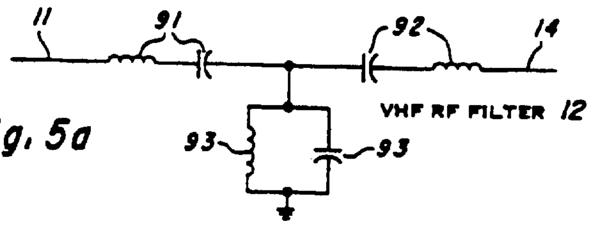


Fig. 5a

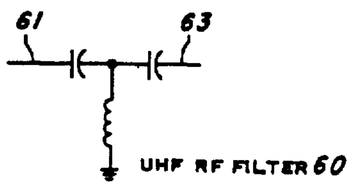


Fig. 5b

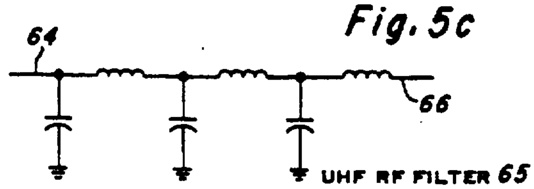


Fig. 5c

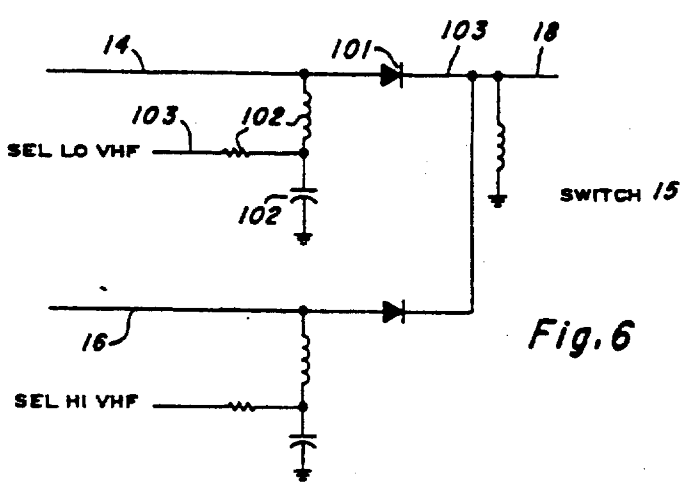


Fig. 6

Fig. 7

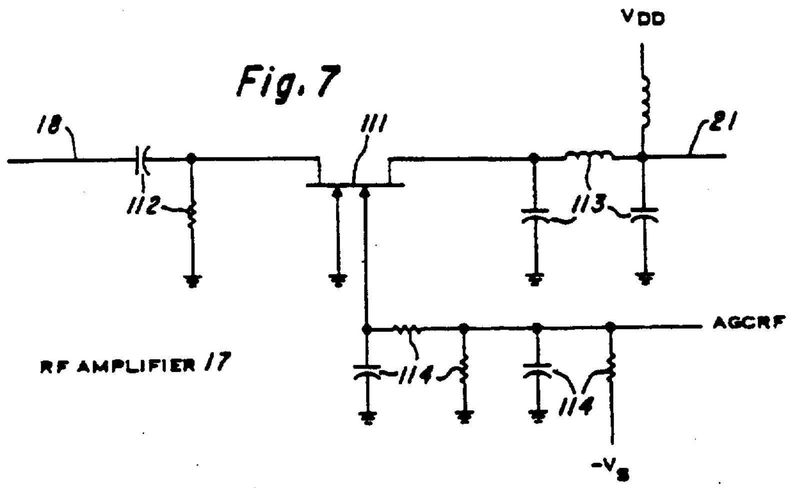
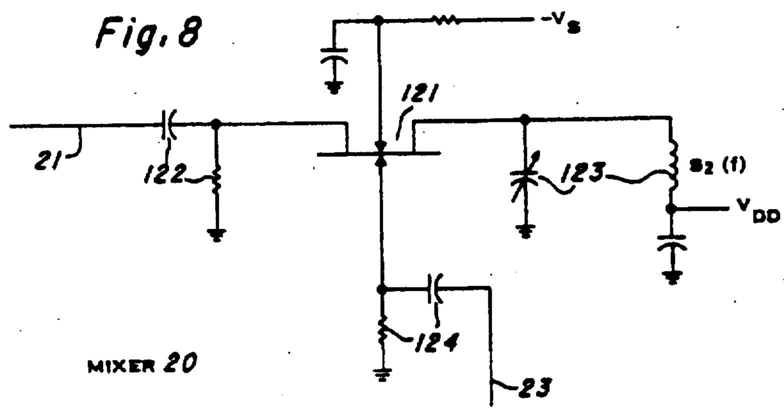


Fig. 8



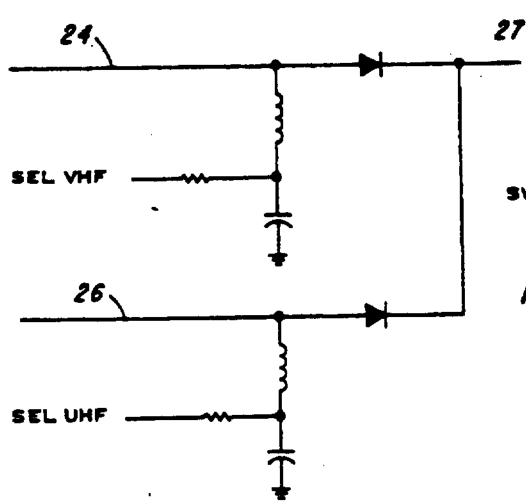


Fig. 10

Fig. 9a

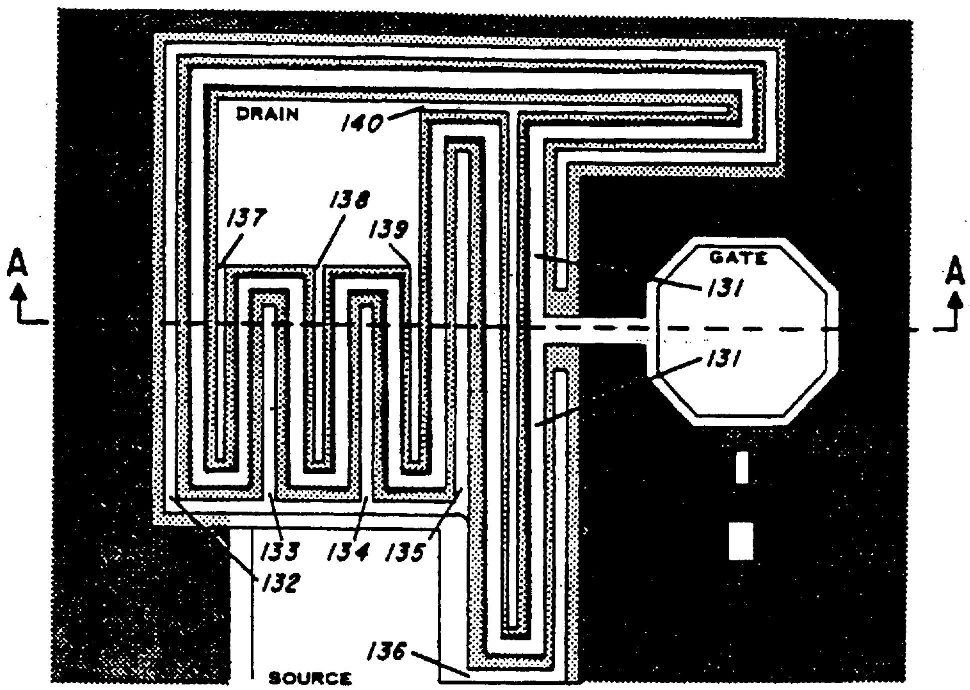


Fig. 9b

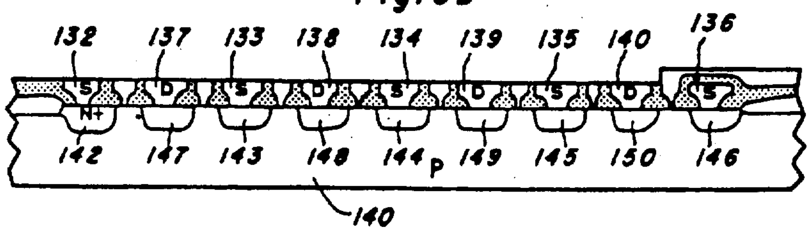


Fig. 9c

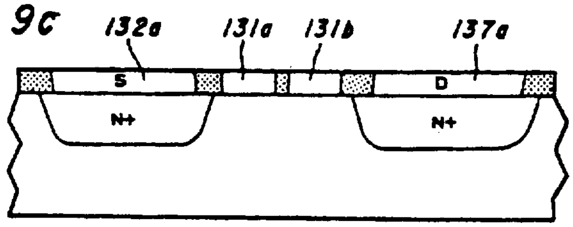


Fig. 9d

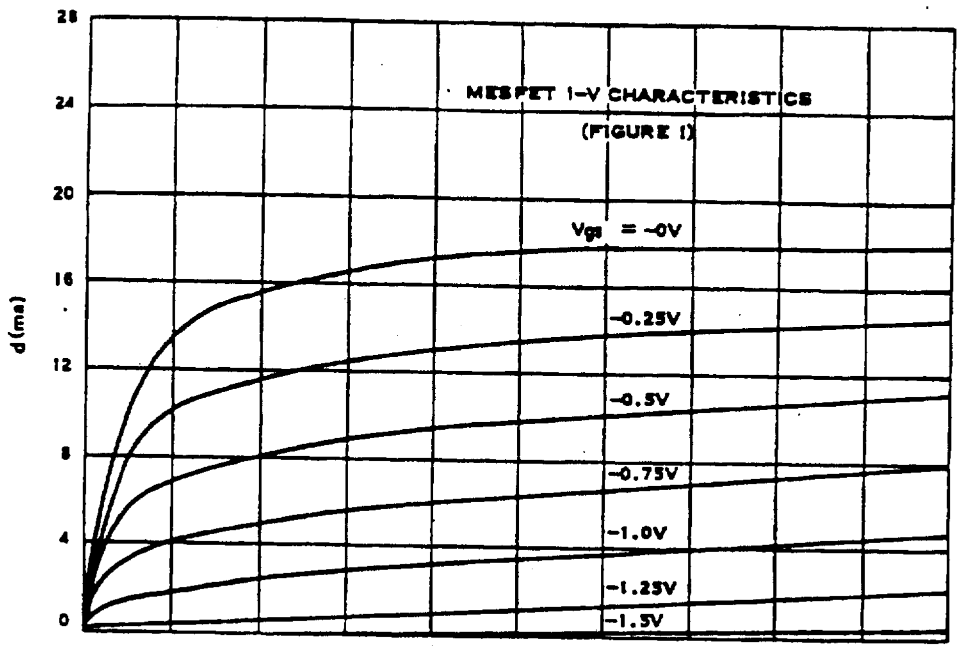

MESFET I–V CHARACTERISTICS
(FIGURE I)
Vgs = –0V
–0.25V
–0.5V
–0.75V
–1.0V
–1.25V
–1.5V
d (ma)
28
24
20
16
12
8
4
0

Fig. 9e

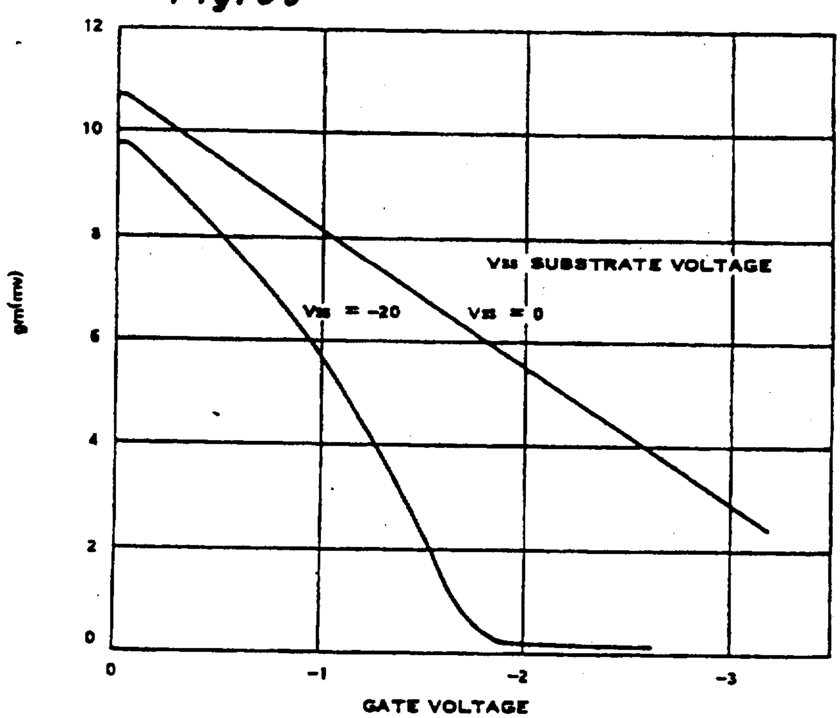

VSS SUBSTRATE VOLTAGE
VSS = –20
VSS = 0
gm(mv)
12
10
8
6
4
2
0
0
–1
–2
–3
GATE VOLTAGE

Fig. 11a

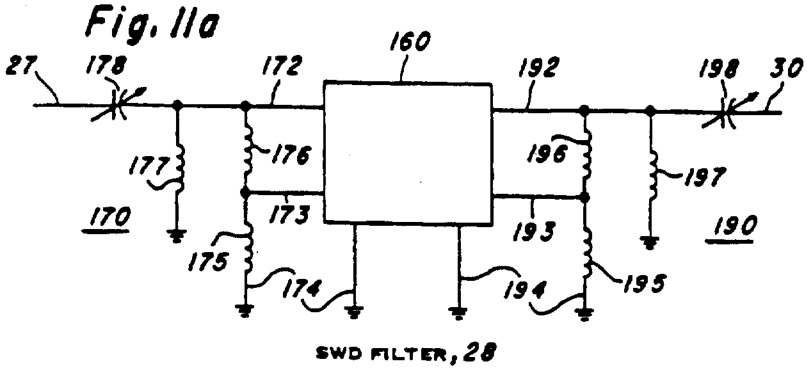
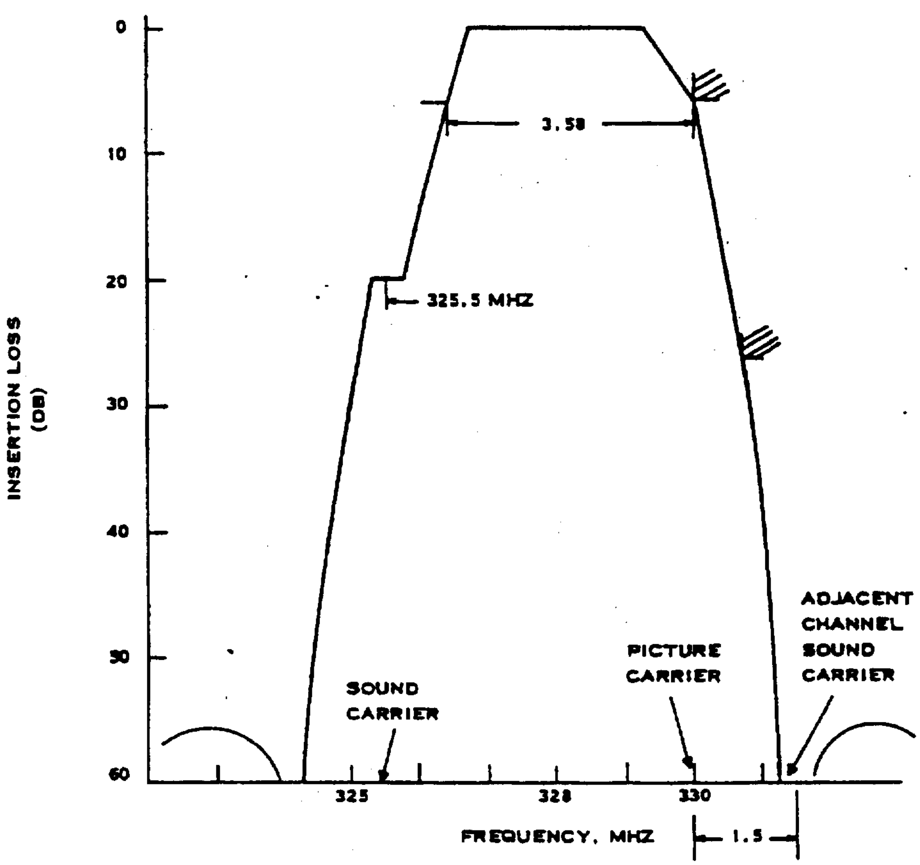


Fig. 11b



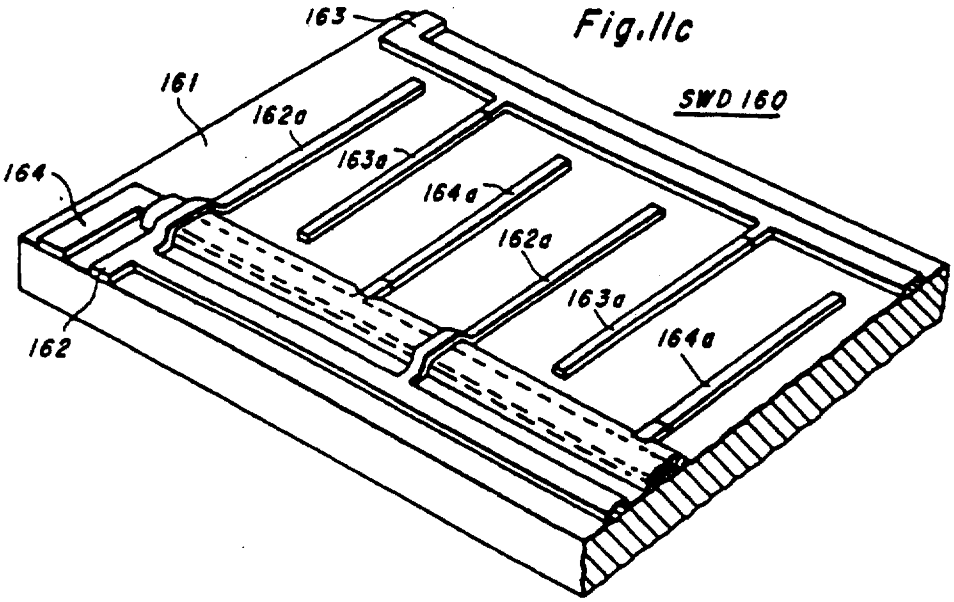


Fig. 11c

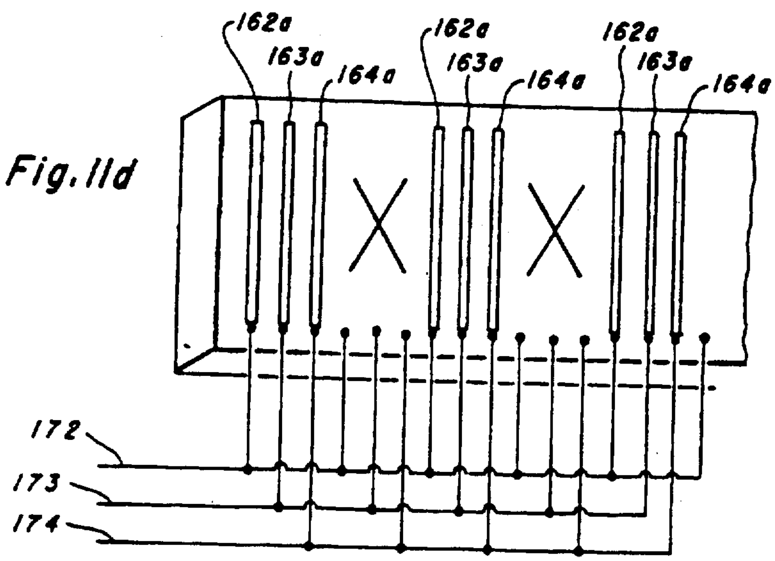


Fig. 11d

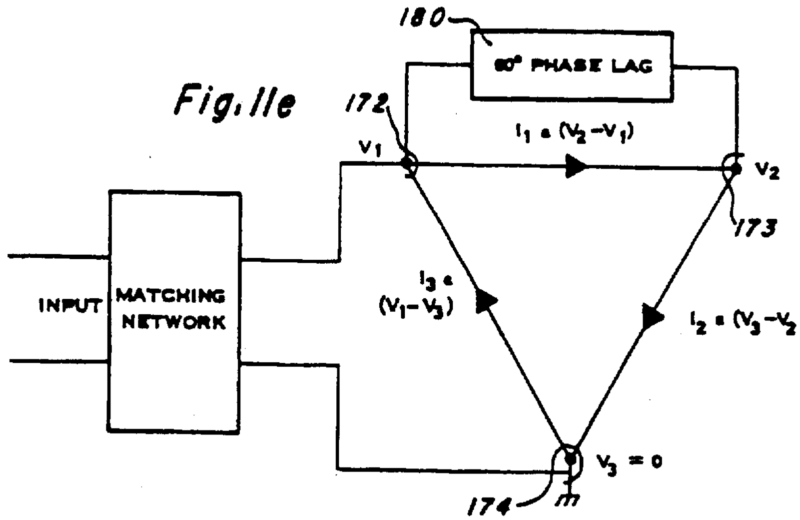


Fig. 11e

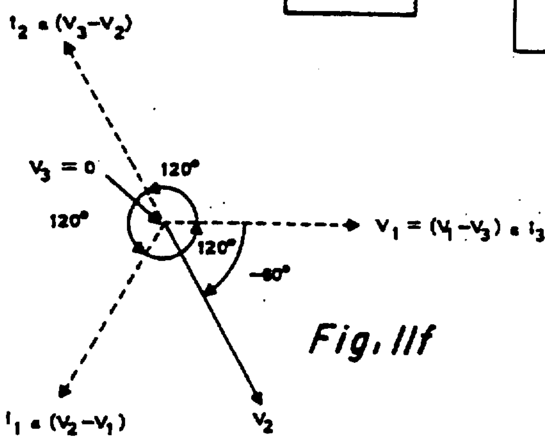


Fig. 11f

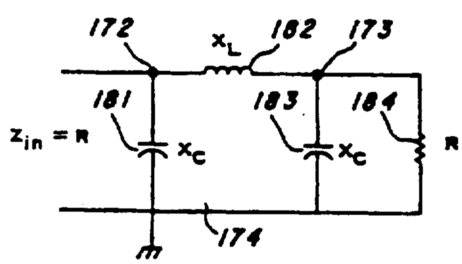


Fig. 11g

$$|X_L| = R \sin \theta$$

$$|X_C| = \frac{R \sin \theta}{(1 - \cos \theta)}$$

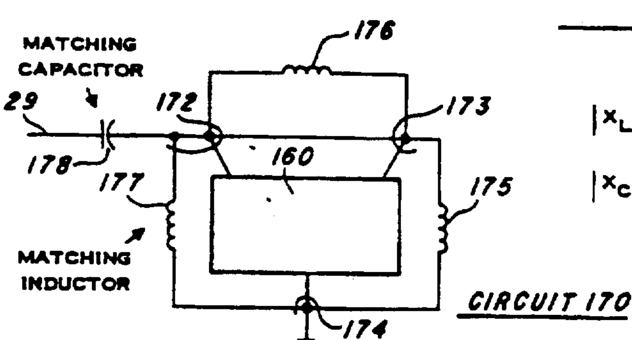
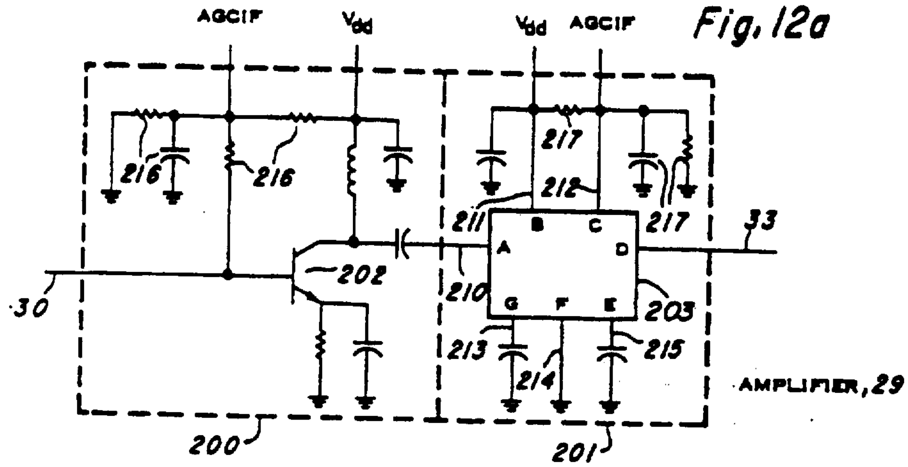


Fig. 11h

Fig. 12a



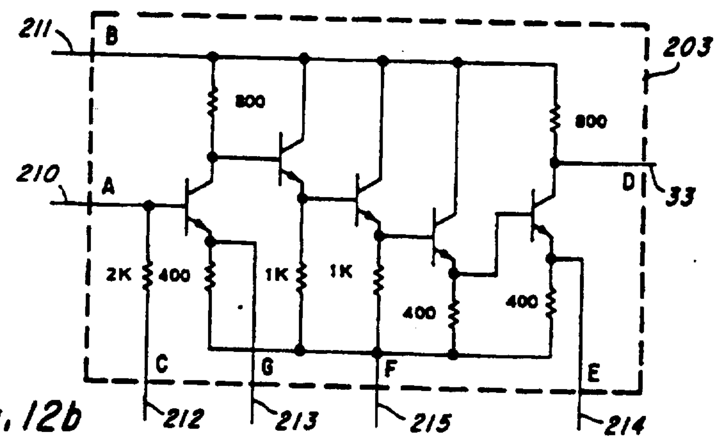


Fig. 12b

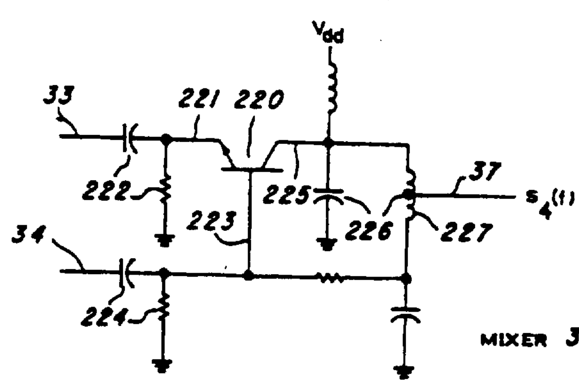


Fig. 13

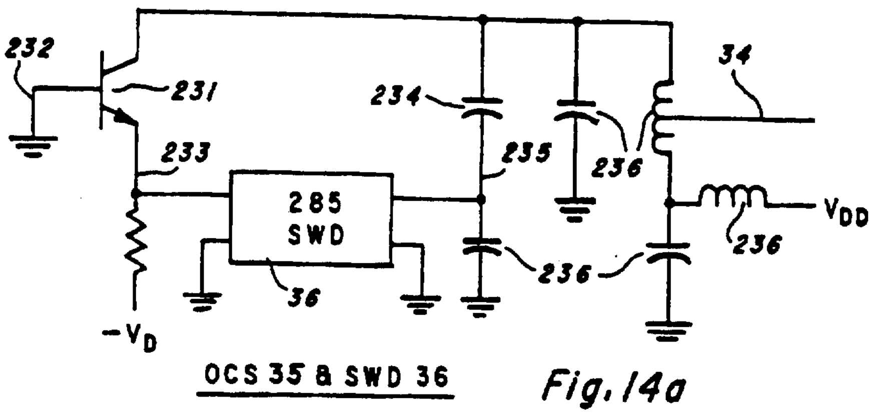


Fig. 14a

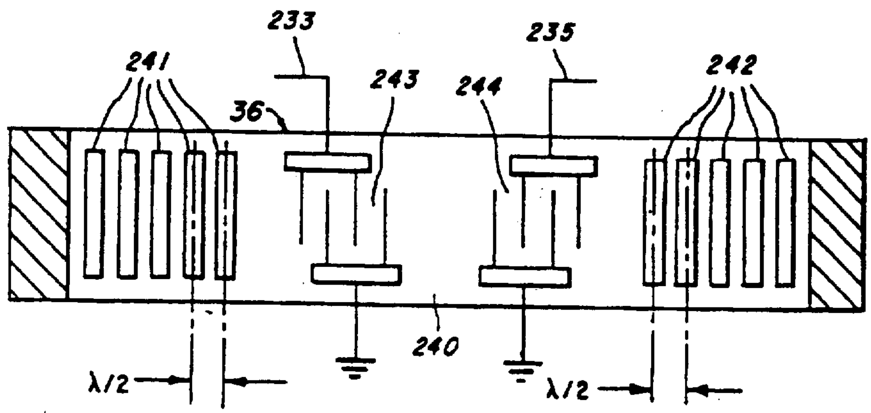


Fig. 14b

Fig. 15

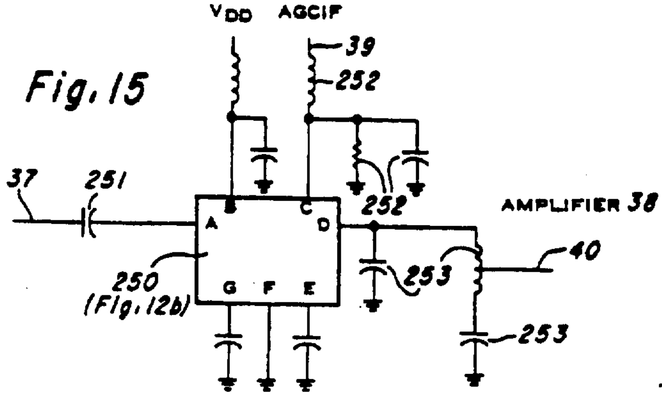


Fig. 16

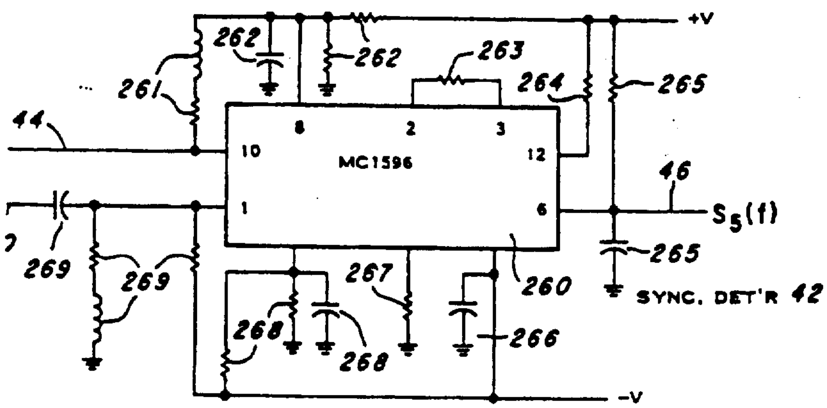


Fig. 17

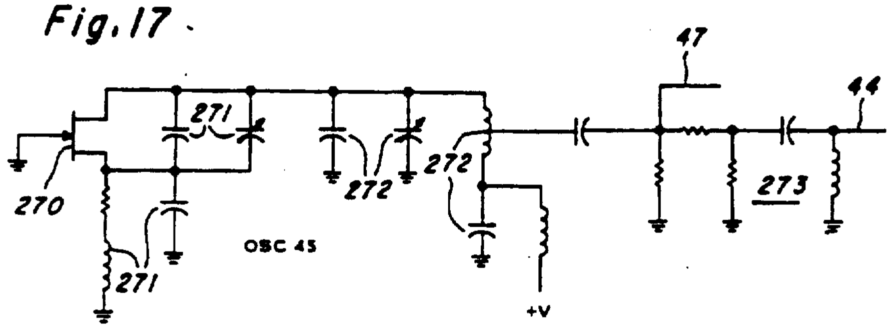
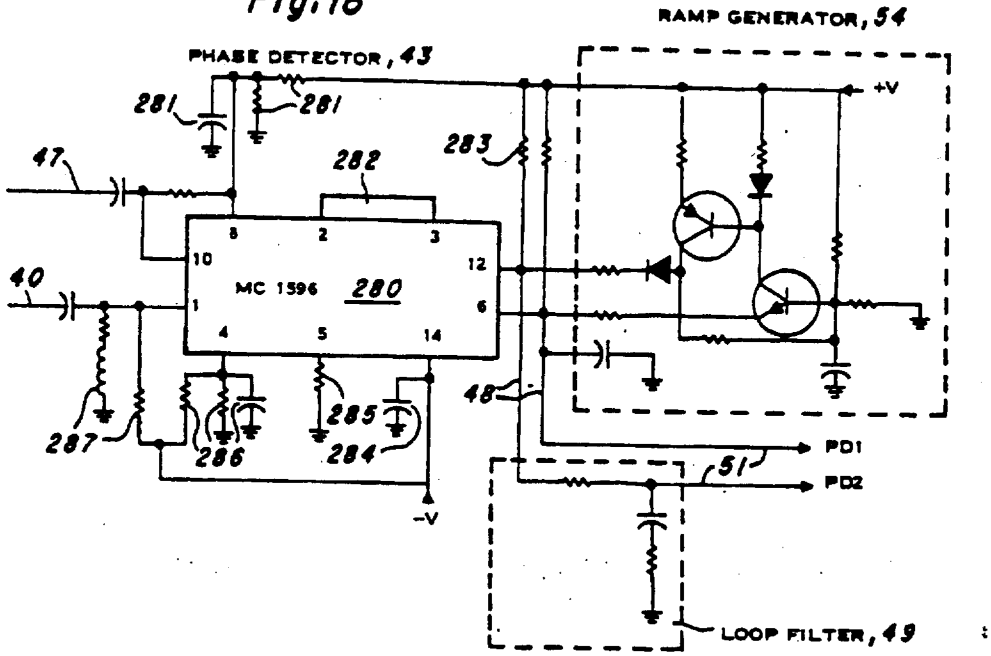


Fig. 18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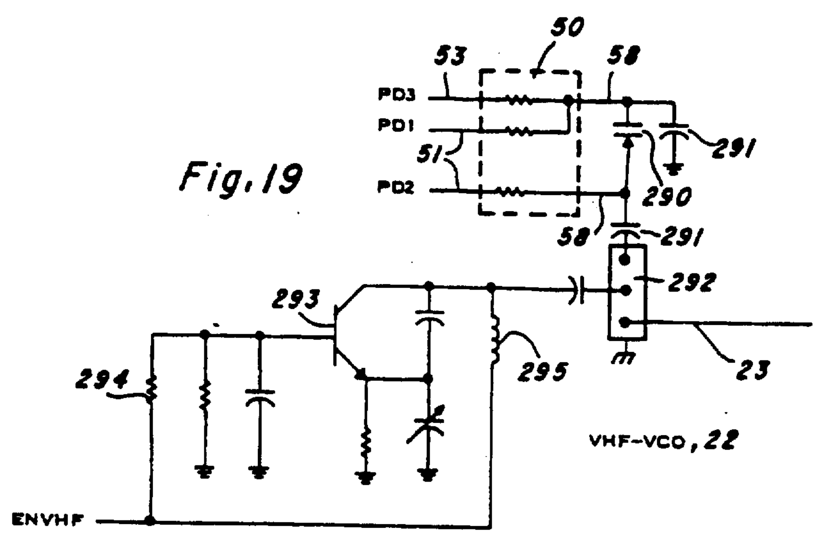
PHASE DETECTOR, 43
281
281
+V
283
282
47
MC 1596
280
8
2
3
10
12
1
6
4
5
14
40
48
285
287
286
284
PD1
51
PD2
-V
LOOP FILTER, 49

Fig. 19


50
53
58
PD3
PD1
51
PD2
291
290
58
291
292
293
294
295
23
VHF-VCO, 22
ENVHF

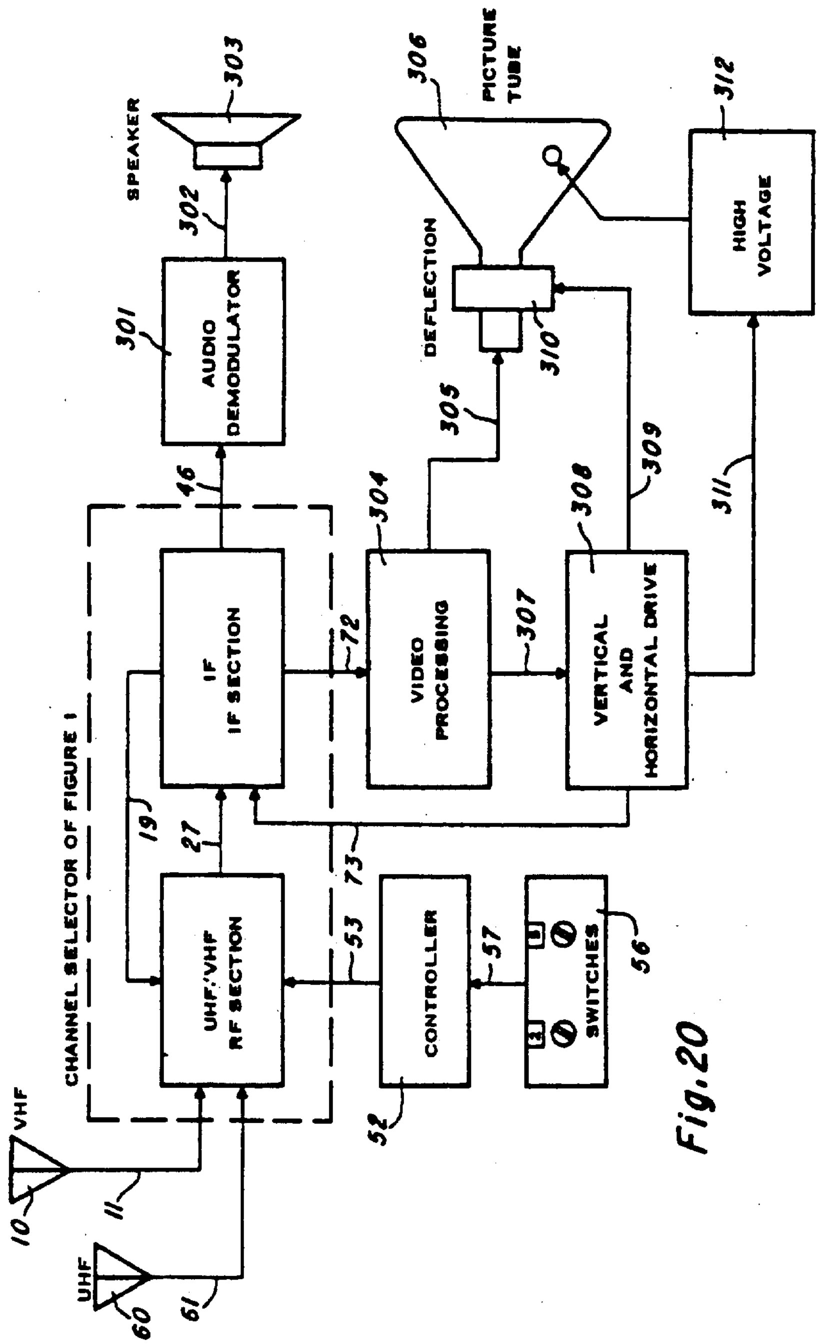

SPEAKER
303
302
AUDIO DEMODULATOR
301
PICTURE TUBE
306
DEFLECTION
310
305
HIGH VOLTAGE
312
311
309
308
304
VIDEO PROCESSING
307
VERTICAL AND HORIZONTAL DRIVE
46
IF
IF SECTION
72
CHANNEL SELECTOR OF FIGURE 1
19
27
73
UHF/VHF RF SECTION
53
CONTROLLER
57
SWITCHES
56
52
VHF
10
11
UHF
60
61

Fig. 20

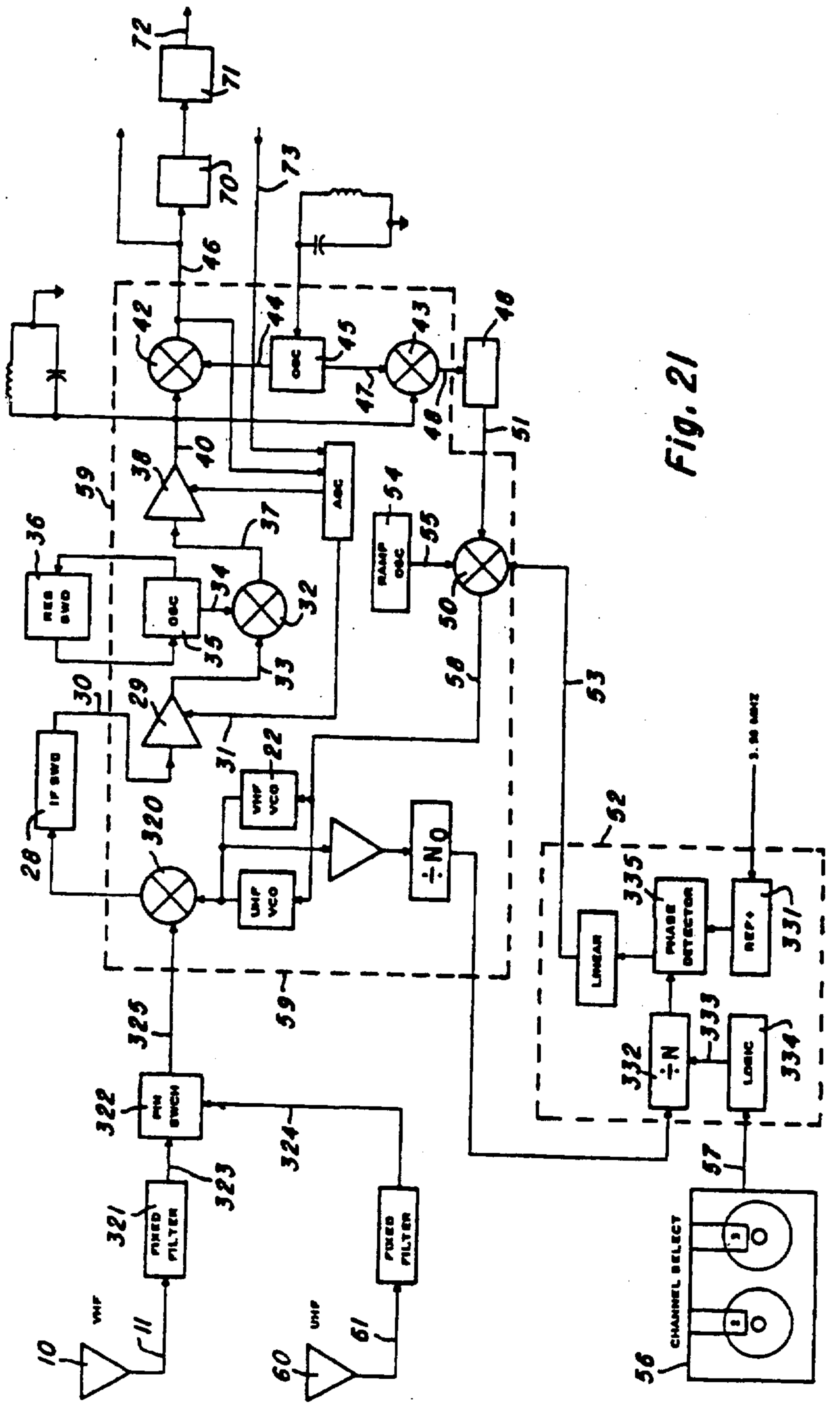


Fig. 21

CHANNEL SELECTION FOR A TELEVISION RECEIVER HAVING LOW-GAIN HIGH FREQUENCY RF-IF SECTION

BACKGROUND OF THE INVENTION

This invention relates to electronic devices for receiving a plurality of radiated electromagnetic signals, filtering a selectable channel of frequencies from the received signals, and demodulating the signals of the selected filtered channel. More particularly, the invention relates to television receivers.

Television receivers of the prior art include a radio frequency (RF) section and an intermediate frequency (IF) section. The RF section includes RF filters which are tuned to coarsely filter a band of channels centered about a manually selected channel. The output of the RF filter couples to the input of an RF amplifier. Typically, total gain through the RF section is at least 20 to 30 dB. This gain increases the amplitude of signals within the selected channel and additionally makes the noise figure of the system essentially independent of subsequent elements in the receiver. The output of the RF amplifier couples to one input of a mixer, while a second input of the mixer receives mixing signals of a selectable frequency. The selectable frequency is generated such that the selected channel is frequency shifted to approximately 45 MHZ. The output of the mixer couples to a channel selection filter which provides a relatively high impedance path for frequencies outside of the selected channel, and a relatively low impedance path for signals inside the selected channel. Signals at the output of the channel selection filter are therefore primarily comprised of frequencies within the selected channel.

Each television channel contains audio information, video information, and frame synchronizing information. The output of the channel selection filter couples to an audio demodulator which separates the audio information from the selected channel; and the output of the audio demodulator couples to a speaker which generates audible sounds. Similarly, the output of the channel selection filter couples to a video processing unit which separates the video and frame synchronizing information from the selected channel; and the output of the video processor couples to a picture tube which converts the video and frame synchronizing information to pictures.

As described above, prior art television receivers insert at least 20 to 30 dB of gain in the RF section to achieve a low system noise figure. The system noise figure equals

$$NF_1+(NF_2-1/G_1)+(NF_2-1/G_1G_2)+(NF_2-1/G_1G_2G_3)+\ldots$$

wherein NF_i and G_i are the noise figure and gain of the ith functional block in the system. Thus, inserting a large gain in the first functional blocks (i.e. the RF section) lowers the system noise figure by making it independent of the noise figure of subsequent circuitry.

However, a high gain RF section generates output terms which are proportional to its inputs cubed or raised to higher order odd powers. This is because the gain occurs before channel filtering, and signals in an undesired channel invariably become so large as to exceed the dynamic range of the circuits. In a television receiver, these odd power terms may generate interfering signals in the desire channel. For example, such interfering signals are generated when signals are present in channels on one side of the selected channel which are one and two channels removed from the selected channel. This phenomena is known as intermodulation distortion. Similarly, the cubic terms and higher order odd power terms generate interfering signals in the desired channel when a carrier with amplitude modulation is present in any one of the undesired channels. This phenomena is known as cross-modulation distortion. The frequencies which are generated in a desired channel as a result of intermodulation distortion or cross-modulation distortion cannot be separated from the information signals lying therein. Thus, as the magnitude of the interfering frequencies increases, perceptible picture distortion or sound distortion occurs.

By comparison, the disclosed invention has a unique architecture which achieves a low noise figure, low intermodulation distortion and low cross-modulation distortion simultaneously. In the preferred embodiment, the disclosed invention has a maximum gain in the channel selecting section which is less than 10 dB; that utilizes circuit elements therein which individually have low noise figures. The 10 dB gain in combination with the individual low noise figures yield a low system noise figure, while the low gain allows RF and IF circuit elements to operate within their dynamic range without generating high odd order output terms.

The architecture of the disclosed invention is also novel in that it has two IF frequencies about which the system gain is selectively distributed. In one preferred embodiment, a first mixer shifts the selected channel to about 330 MHz. The output of this mixer couples to the channel selection filter; and the output of the channel selection filter couples to a second mixer which frequency shifts the filtered selected channel to a second IF frequency of approximately 45 MHz. Most of the gain of the system (approximately 60 dB) is inserted at the lower IF frequency after the channel selection filter. Gain at the high IF frequency is small as pointed out above. As a result, several advantages of a high IF system - such as simplified filtering of image frequencies-are obtained, while feedback inherent in a high gain-high IF section is avoided.

The disclosed system is also simpler and potentially less expensive than prior art television receivers in that the low gain RF section makes feasible integration of a major portion of the receiver on a single semiconductor chip. Incorporated in the disclosure are two differently organized RF-IF sections which are suitable for semiconductor chip integration.

The intermodulation distortion and cross-modulation distortion in a preferred embodiment are further reduced by utilizing a MESFET mixer (i.e. a mixer including a metal semiconductor field effect transistor) in the low gain RF section. The MESFET mixer has almost perfect square law characteristics when operated at a low gain, and consequently the device introduces extremely small distortion into the system. In particular, the device can handle an interfering signal level at greater than +6 dBm on its output with less than 1% cross-modulation distortion and −40 dB intermodulation distortion products. Prior art television receivers, by comparison, utilize mixers comprised of bipolar or MOS transistors, diodes, or vacuum tubes, all of which generate third order products or higher odd order products to a greater degree than the disclosed MESFET mixer.

Accordingly, it is one object of the invention to provide an improved device for receiving signals in selected frequency channels from a plurality of non-overlapping frequency channels.

Another object of the invention is to provide a television receiver having improved reception.

Another object of the invention is to provide a television receiver having a low noise figure, low intermodulation distortion, and low cross-modulation distortion simultaneously.

Another object of the invention is to provide a television receiver having an RF section with a gain just large enough to achieve a desired system noise figure.

Another object of the invention is to provide a television receiver having a maximum gain before the channel selection filtering of approximately 10 dB.

Another object of the invention is to provide a television receiver having a low gain-high frequency RF-IF section.

Another object of the invention is to provide a television receiver having a low gain-high frequency RF-IF section in combination with a high gain-low frequency IF section.

Another object of the invention is to provide a television receiver suitable for integration on a single semiconductor chip.

Still another object of the invention is to provide a television receiver having a low gain MESFET mixer for frequency shifting the selected channel to a predetermined high IF.

SUMMARY OF THE INVENTION

These and other objectives are accomplished in accordance with the invention by a channel selector having a low-gain high-frequency RF-IF section. The channel selector includes a frequency spectrum filter in the RF section, a mixer, and a channel selecting filter. The frequency spectrum filter is coupled to the antenna of the television receiver and filters a frequency spectrum of radio frequency signals as received from the antenna, such as a plurality of television channels. The mixer has an input coupled to the frequency spectrum filter and frequency shifts selected channels of the frequency spectrum to a predetermined relatively high intermediate frequency. One embodiment utilizes a low-gain MESFET mixer. The channel selecting filter has an input coupled to the mixer and filters the channel at the predetermined intermediate frequency. Total gain from the antenna through the channel selecting filter is no larger than necessary to obtain a desired system noise figure. In one embodiment, the total gain is less than 10 dB. The IF section includes a second mixer coupled to the output of the channel selecting filter for frequency shifting the filtered selected channel to a lower second intermediate frequency. An IF amplifier having a relatively high gain is connected to the output of the second mixer and adds gain to the filtered selected channel at the lower second intermediate frequency, the added gain being substantially in excess of the total gain provided before the channel selecting filter.

BRIEF DESCRIPTION OF THE DRAWINGS

The novel features believed characteristic of the invention are set forth in the appended claims; the invention itself, however, as well as other features and advantages thereof, may best be understood by referring to the following detailed description of particular embodiments when read in reference to the accompanying drawings, wherein:

FIG. 1 is a block diagram illustrating a channel selector constructed according to the invention.

FIGS. 2a–2e are a series of frequency diagrams illustrating signals at selected points on the channel selector of FIG. 1.

FIG. 3 is a timing diagram illustrating the operation of a phase locked loop within the channel selector of FIG. 1.

FIGS. 4a and 4b are graphs respectively illustrating the amplitude of in-band and out-of-band signals at various points in the channel selector of FIG. 1.

FIGS. 5a, 5b and 5c are detailed circuit diagrams of RF filters within the channel selector of FIG. 1.

FIG. 6 is a detailed circuit diagram of a two-by-one switch within the channel selector of FIG. 1.

FIG. 7 is a detailed circuit diagram of an RF amplifier within the channel selector of FIG. 1.

FIG. 8 is a detailed circuit diagram of a MESFET mixer within the channel selector of FIG. 1.

FIG. 9a is a plan view of a MESFET transistor suitable for application in the MESFET mixer of FIG. 8.

FIG. 9b is an enlarged cross-sectional view taken along the line A—A of FIG. 9a.

FIG. 9c is a greatly enlarged cross-sectional view of a dual gate MESFET device showing one source-drain pair thereof and which comprises another MESFET transistor suitable for application in the MESFET amplifier of FIG. 7.

FIGS. 9d and 9e are graphs illustrating the operational characteristics of the MESFET transistor shown in FIG. 9a.

FIG. 10 is a circuit diagram of a second two-by-one switch within the channel selector of FIG. 1.

FIG. 11a is a circuit diagram of a surface wave device acoustic filter within the channel selector of FIG. 1.

FIG. 11b is a graph illustrating the magnitude-frequency characteristic of the surface wave device included in the filter of FIG. 11a.

FIG. 11c is a perspective view illustrating one embodiment of the surface wave device included in the filter of FIG. 11a.

FIG. 11d is a partially perspective, diagrammatic view illustrating the surface wave device included in the filter of FIG. 11a where the impulse response of the surface wave device has been shaped by electrode finger withdrawal.

FIG. 11e is a diagram of a phase shifting circuit for use in the input circuitry of the filter of FIG. 11a.

FIG. 11f is a phase diagram illustrating how the phase shifting circuit of FIG. 11e can produce a 120° phase shift.

FIG. 11g is a diagram of a circuit for producing a 60° phase lag.

FIG. 11h is a circuit diagram of the input circuitry included in the filter of FIG. 11a. *FIGS. 12a–12b* are detailed circuit diagrams of a linear amplifier within the channel selector of FIG. 1.

FIG. 13 is a detailed circuit diagram of a mixer within the channel selector of FIG. 1.

FIG. 14a is a detailed circuit diagram of a surface wave device oscillator within the channel selector of FIG. 1.

FIG. 14b is a top plan view illustrating the surface wave device included in the oscillator of FIG. 14a.

FIG. 15 is a detailed circuit diagram of a high gain linear amplifier within the channel selector of FIG. 1.

FIG. 16 is a detailed circuit diagram of a synchronous detector within the channel selector of FIG. 1.

FIG. 17 is a detailed circuit diagram of a L-C oscillator within the channel selector of FIG. 1.

FIG. 18 is a detailed circuit diagram of a phase detector, a ramp-generator, and a loop filter within the channel selector of FIG. 1.

FIG. 19 is a detailed circuit diagram of a voltage controlled oscillator within the channel selector of FIG. 1.

FIG. 20 is a block diagram of a television receiver which includes the channel selector of FIG. 1.

FIG. 21 is a block diagram of an alternative embodiment for the channel selector of FIG. 1.

DETAILED DESCRIPTION OF SPECIFIC EMBODIMENTS

Referring to FIG. 1, the channel selecting portion of a television receiver constructed according to the invention is illustrated in block diagram form. The selector includes a VHF antenna 10 having an output coupled via a lead 11 to a fixed bandpass filter 12 and to another fixed bandpass filter 13. Signals on lead 11 are herein designated $S_1(f)$. Filter 12 has a center frequency of 69 MHZ and a 3 dB bandwidth of 34.6 MHZ to thereby permit passage of the low VHF television channels. Similarly, filter 13 has a center frequency of 193 MHZ and a 3 dB bandwidth of 44.5 MHZ to provide filtering of the high VHF television channels. The output of filter 12 is coupled via a lead 14 to one of the inputs of a 2×1 switch 15, while the output of filter 13 is coupled via a lead 16 to a second input of switch 15. Switch 15 operates to select either signals from filter 12 or filter 13. The filter-switch combination inserts approximately a 1 dB loss on signals in the passbands.

The output of switch 15 is coupled to an RF amplifier 17 via lead 18. Amplifier 17 has a variable gain which is controlled by signals on an AGC line 19. The maximum gain of amplifier 17 is approximately 4 dB. The noise figure of the amplifier is approximately 3 dB.

The output of amplifier 17 couples to one input of a MESFET mixer 20 via a lead 21. Mixer 20 has a fixed gain of approximately 4 dB, and a noise figure of approximately 8 dB. A second input of mixer 20 is coupled to a VHF voltage controlled oscillator 22 via a lead 23. Oscillator 22 generates local oscillator (LO) signals on lead 23 of a selectable frequency in the range 385–541 MHZ. In response thereto, mixer 20 frequency shifts the RF signals on lead 21 to a new IF frequency range. The frequency of the LO signals on lead 23 is selected such that the channel to be received is frequency shifted to a predetermined high IF of between 300 MHZ to 400 MHZ. In one embodiment, this predetermined high IF is approximately 330 MHZ. The frequency shifted signals are generated on a lead 24 and are designated $S_2(f)$.

Lead 24 couples to one input of a 2×1 switch 25, while a second input of switch 25 is coupled via a lead 26 to receive frequency shifted UHF television channels. Switch 25 is identical in construction to switch 15. The circuitry for frequency shifting the UHF television channels, which is labeled as UHF RF SECTION, is similar in construction to the VHF-RF section, and is described in detail infra.

The output of switch 25 is coupled to a surface wave device (SWD) filter 28 via a lead 27. Filter 28 has a passband which is shaped to pass only one of the television channels on lead 27. In particular, filter 28 passes the television channel at the predetermined high IF.

The fixed RF filters in combination with the SWD filter provide essentially all the filtering in the system. In a preferred embodiment, filter 28 is a three phase unidirectional filter which has a low insertion loss. Typically, the loss through filter 28 is less than 3.5 dB in the passband. Conversely, out of band signals are greatly attenuated.

The output of filter 28 is coupled to an IF amplifier 29 via a lead 30. The signals on lead 30 are designated herein as $S_3(f)$. Amplifier 29 has a variable gain which is controlled by AGC signals on a lead 31. The maximum gain of amplifier 29 is approximately 30 dB. Thus, amplifier 29 is the first high gain element in the system. To this point, the gain of the system is characterized as being no larger than necessary to obtain the desired system noise figure. Noise figure for amplifier 29 is approximately 4 dB.

The output of amplifier 29 couples to a mixer 32 via a lead 33. A second input of mixer 32 is coupled via a lead 34 to the output of an oscillator 35. Oscillator 35 has a surface wave device resonator 36 as its frequency controlling element. In a preferred embodiment, resonator 36 and oscillator 35 operate to generate a mixing signal on lead 34 having a frequency of 285 MHZ. The 285 MHZ signal on lead 34 is mixed with the signal on lead 33 to thereby generate signals $S_4(f)$ on the output of mixer 32. Signal $S_4(f)$ is similar to signal $S_3(f)$ except that it is shifted down in frequency by 285 MHZ. Thus it has a picture carrier at approximately 45 MHZ.

Additional gain is added to the signal of the selected channel after it has been frequency shifted to the lower IF frequency of 45 MHZ. Mixer 32 adds a fixed gain of +10 dB and the output of mixer 32 couples via a lead 37 to an IF amplifier 38 which has a maximum gain of +50 dB. The gain of amplifier 38 is varied by an AGC signal on a lead 39. By inserting most of the gain at the lower IF frequency, system stability is increased since feedback through parasitic capacitances, radiation, etc. is much less at 45 MHZ than 330 MHZ.

The output of amplifier 38 is coupled via lead 40 to a tank circuit 41, a synchronous detector 42 and the phase detector 43. Tank circuit 41 has a center frequency of approximately 45 MHZ. Synchronous detector 42 has a second input coupled via a lead 44 to an oscillator 45. Oscillator 45 generates clock signals at a fixed frequency of 45 MHZ on lead 44. The signals on lead 44 are in phase with the 45 MHZ picture carrier on lead 40. Detector 42 mixes the signals on leads 40 and 44 to generate output signals $S_5(f)$ on lead 46. Signal $S_5(f)$ contains the selected television channel with the picture carrier at zero Hz and the sound carrier at 4.5 MHZ. Lead 46 is then coupled to conventional television circuitry for separating the sound signal from the picture signals and for reproducing the sound and the picture from these signals respectively in a conventional manner.

The synchronous clock signal on lead 44 is kept in phase with the picture carrier on lead 40 by means of the phase detector 43 operating in conjunction with oscillator 45. Phase detector 43 generates phase detection signals PD2 on lead 48 which indicate the phase difference between the picture carrier on lead 40 and the oscillator signal on lead 47. Signals PD2 maintain a 90° phase difference between the signals on leads 40 and 47. This phase difference is compensated for by oscillator 45 which maintains a compensating 90° phase difference between the oscillator signals on leads 44 and 47.

To complete the phase locked loop, lead 48 couples to a loop filter 49, and the output of filter 49 couples to a summer 50 via leads 51. Summer 50 has a second input coupled to a controller 52 via a lead 53, and a third input coupled to a ramp oscillator 54 via lead 55. Controller 52 and oscillator 54 operate to provide a coarse voltage for selecting one channel from another. Channel selection switches 56 are coupled to controller 52 via leads 57. Switches 56 generate digital signals on leads 57 indicating the selected channel. Controller 52 contains a digital to analog converter which generates a coarse channel selection voltage on lead 53 in response to the digital signals. The signals on lead 53 are summed with the coarse signals on leads 51 and 55 to thereby provide a loop which is phase locked to the picture carrier of the selected channel. The output of summer 50 is coupled to VHF voltage controlled oscillator 22 via leads 58. Oscillator 22 generates LO signals on lead 23 of a frequency ranging from 385 MHZ to 541 MHZ in response to the phase detection signals on leads 58 thereby completing the loop.

The UHF-RF section of the FIG. 1 channel selector begins with a UHF antenna 59. Antenna 59 is coupled to a high pass filter 60 via a lead 61. Filter 60 has a fixed 3 dB cutoff frequency of approximately 380 MHZ. The output of filter 60 couples to the input of an RF amplifier 62 via a lead 63. Amplifier 62 also has an AGC input which is connected to receive AGC signals on lead 19. The maximum gain of amplifier 62 is 4 dB, and its noise figure is also approximately 4 dB. A low pass filter 65 has its input coupled to the output of amplifier 62. Filter 65 has a fixed 3 dB cutoff frequency of approximately 936 MHZ. Thus, filter 65 in combination with filter 60, pass the entire UHF band and reject other frequencies.

A mixer 67 receives the UHF band of signals from filter 65 via a lead 66. Mixer 67 simultaneously receives mixing signals of a selectable frequency from a UHF voltage controlled oscillator 68. These selectable frequencies range from approximately 801 MHZ to 1215 MHZ. The particular frequency at any time instant is generated in response to channel selection switches 56 so as to frequency shift the selected channel to the predetermined high IF.

The overall operation of the above described FIG. 1 structure is as follows. Antennas 10 and 59 receive radiated electromagnetic signals which include the VHF and UHF frequency spectrum. The signals received by antenna 10 are applied to non-tuned filters 12 and 13 which respectively pass the entire low VHF and high VHF frequency spectrum. Similarly, the signals received by antenna 59 are applied to non-tuned filters 60 and 65 which pass the entire UHF frequency spectrum. Switches 15 and 25 select one of these three frequency spectrums in response to logic signals from the channel selection switches 56.

MESFET mixers 20 or 67 then frequency shift the selected frequency spectrum to a predetermined IF in the 300 MHZ-400 MHZ range. Mixer 20 frequency shifts VHF signals, whereas mixer 67 frequency shifts UHF signals. Both mixers have a large dynamic range over which their output is an almost perfect product of their inputs. As a result, the system has improved performance. For example, mixers 20 and 67 can handle $>+6$ dBm interfering signal levels on their output with less than 1% cross-modulation distortion.

Switch 25 then couples one of the mixers to the input of SWD filter 28. Filter 28 greatly attenuates signals outside the selected channel. In particular, the stopband of filter 28 is notched such that the lower adjacent sound carrier and upper adjacent picture carrier are attenuated by more than 65 dB. All other out of band signals are attenuated by at least 55 dB. Conversely, in band insertion loss of filter 28 is only 3.5 dB.

Output signals from filter 28 are sent to amplifier 29 which is the first high gain component of the system. Amplifier 29 has a maximum gain of 30 dB. In comparison, the total gain of the system before amplifier 29 is less than 10 dB. As a result of the low RF-IF gain, intermodulation distortion and cross modulation distortion is decreased. Further, system stability is improved since high frequency feedback due to parasitic capacitance, radiation, etc. is avoided. At the same time, a low system noise figure is achieved by circuit elements which individually have a low noise figure in combination with a sufficiently high gain.

The low gain RF-IF section also makes feasible integration of all of the channel selectors, except the fixed filters, on a single semiconductor chip. An outline of such a chip is designated by the dashed line 69 in FIG. 1. The process for fabricating the chip includes a combination of steps presently well known for constructing MESFET devices and bipolar devices. A second chip is utilized to construct SWD filter 28 and SWD resonator 36. The fixed filters 12, 13, 60 and 65 are constructed of discrete components.

Signals in the selected channel at the output of amplifier 29 are frequency shifted to 45 MHZ by mixer 32. Then they are further amplified to -10 dBm by amplifier 38. The selected channel video signals are then reduced to baseband by synchronous detector 42; while the sound carrier of the selected channel is shifted to 4.5 MHZ. A 4.5 MHZ trap 70 removes the sound from the video signals, and a low pass filter 71 removes all signals except the video of the selected channel. The video signals at the output of filter 71 are sent via lead 72 to video processing circuitry, while the audio signals at 4.5 MHZ on lead 46 are sent to audio processing circuitry. This audio-video circuitry is described infra in conjunction with FIG. 20.

Referring now to FIGS. 2a–2e, there is illustrated a set of frequency diagrams of signals $S_1(f)$–$S_5(f)$. In FIG. 2a, the low VHF band is shown generally at 75a, the high VHF band is shown generally at 75b, and the UHF band is shown at 75c. Each of the bands is comprised of a plurality of channels; and each channel has a frequency spectrum which is assigned as illustrated in detail at 76. The frequency allocation and type of modulation of signals within each channel is a well known standard that is fixed by the FCC. FIG. 2b is an exemplary frequency diagram of signal $S_2(f)$. In the example illustrated, signal $S_2(f)$ contains channels in the low VHF spectrum. The selected channel is near 330 MHZ. The frequency of the LO signal on lead 23 minus the frequency of the picture carrier from the selected channel equals 330 MHZ. Since the mixing frequency is higher than the picture carrier, the frequency spectrum at the output of mixer 20 is inverted from the input frequency spectrum as shown at 77.

FIG. 2c illustrates the frequency spectrum of signal $S_3(f)$. Signal $S_3(f)$ is the output of the channel selecting filter 28. Thus it basically contains only frequencies within the selected channel. Signal $S_3(f)$ is then amplified and frequency shifted down by 285 MHZ. The result is signal $S_4(f)$ as illustrated in FIG. 2d.

Signal $S_4(f)$ is further amplified, and then synchronously detected by detector 42. These operations pro-

duce signal $S_5(f)$ as illustrated in FIG. 2e. Note that the mixing action of synchronous detector 42 again inverts the frequency spectrum of the selected channel. Thus, the picture carrier of the selected channel is at 0 HZ, and the sound carrier of the selected channel is at 4.5 MHZ as shown at 78.

FIG. 3 is a timing diagram illustrating the operation of the channel selecting phase locked loop of FIG. 1. As shown therein, signal PD4 on lead 58 is comprised of components PD1–PD3. Signal PD3 constitutes a coarse channel selection voltage which is produced by controller 52. Signals PD2 and PD1 provide the fine tuning for the phase locked loop. In the example of FIG. 3, one particular channel is selected during a first time interval ΔT1, while another channel is selected during a time interval ΔT2. Signal PD3 provides the coarse voltage for channel selection. Signal PD1 compensates for any DC offset between signal PD3 and the desired voltage level, and signal PD2 provides a dynamic correction voltage to compensate for instantaneous phase or frequency differences between the signals on leads 40 and 44.

The magnitude of the signals of FIGS. 2a–2e are illustrated in FIGS. 4a and 4b. FIG. 4a illustrates the magnitude of signals in the desired channel, whereas FIG. 4b illustrates the magnitude of interfering signals in a channel that is two channels removed. Curves 81–84 illustrate the magnitude of signals in the desired channel at various points in the system when the incoming signal strength is 0 dBM, −35 dBM, −55 dBM, and −85 dBM respectively. As these curves illustrate, gain is first added by amplifier 38, then by amplifier 29, and finally by amplifier 17 as the input signal strength in the selected channel decreases. In particular, the RF section has no gain unless the input signal strength in the selected channel is less than −55 dBm. And the RF section reaches its maximum gain of +3 dBm when the input signal strength of the selected channel is between −55 dBm and −85 dBm.

As previously pointed out, the low gain RF section in combination with the MESFET·RF amplifiers and MESFET RF mixer provides a system having superior channel discrimation capability. This is exemplified in FIG. 4b by the relative signal strength of the selected and unselected channels. For example, curve 85 of FIG. 4b illustrates the case where the input signal strength of the selected channel is −55 dBm and input signal strength two channels removed is +1 dBm. Similarly, curve 86 illustrates the case where the input signal strength in the selected channel is −35 dBm and the input signal strength two channels removed is +2 dBm. The most stringent requirement for the receiver is when the first RF amplifier requires gain (i.e., when the signal strength of the desired channel is less than −55 dBm). This is because the added gain increases nonlinearities in the RF section and aggravates the cross modulation and intermodulation distortion. Thus, curve 85 illustrates the most stringent condition for the system. Under the conditions of curve 85, RF amplifier 17 and mixer 20 must have cross modulation distortion and intermodulation distortion of less than 1%. This requirement is met, since the MESFET designs herein described have less than 1% cross modulation distortion and intermodulation distortion when their output signal level is less than +6 dBm.

After the selected channel has been shifted to 330 MHZ, signals two channels removed from the selected channel are greatly attenuated. Mixer 20 has a tuned

output which attenuates two channel removed signals by −4 DB. SWD filter 28 attenuates all out of band signals by at least −53 DB. At the same time, mixer 20 adds 4 DB of gain to the desired signal, and filter 28 inserts only 3.5 DB of loss to signals in the selected channel.

The details of the various blocks of FIG. 1 will now be described in conjunction with FIGS. 5–19. Referring first to FIG. 5a, a circuit diagram of bandpass filter 12 is therein illustrated. Basically, filter 12 is comprised of two series resonant LC circuits 91 and 92 and one parallel resonant LC circuit 93. Filter 13 is similarly structured.

A primary function of RF filters 12 and 13 is to pass one band of channels while rejecting image frequencies by an amount sufficient to eliminate perceptible picture interference. Since this system uses a 330 MHZ IF, all image frequencies are 660 MHZ above the frequency of the desired channel. Thus, the low VHF images are rejected by the three pole bandpass filter of FIG. 5a by greater than 80 DB. In the system of FIG. 1, just perceptible picture interference occurs when image frequency signals are less than 36 DB below the level of the desired signal at the filter output. Thus, for example, the filter of FIG. 5a provides adequate image frequency rejection when the desired picture signal level is at −55 DBM and the image frequency level is at −11 DBM. Image frequencies for the low VHF and the high VHF spectrum lie within the UHF spectrum, and the level of UHF signal at the input of the VHF antenna can normally be expected to be less than −11 DBM.

FIGS. 5b and 5c are circuit diagrams of a high pass filter and a low pass filter suitable for use by the UHF radio frequency section of FIG. 1. All of the UHF images fall outside of the TV band. The first image falls at approximately 1130 MHZ and the last image falls at approximately 1545 MHZ. These frequencies are allocated for aircraft navigation, with the TACAN system having the highest power output. But TACAN power output is only 5 KW with a 1.8×10^{-3} to 1 duty cycle. And thus the TACAN signal is 50 DB lower than that transmitted by a 1 MW TV transmitter. Therefore, the filters of FIGS. 5b and 5c provide adequate rejection of image signals for the UHF band.

FIG. 6 is a detailed circuit diagram of switch 15. One portion is basically comprised of a diode 101 connecting leads 14 and 18, and an RLC bias network 102 having a control input 103. A DC voltage control signal SELLOVHF is applied to lead 103 for selectively turning diode 101 on or off to thereby select or deselect the low VHF signals on lead 14. Another portion identical to the one described above couples lead 18 to lead 16, and is utilized to select and deselect the high VHF channels.

Referring next to FIGS. 7 and 8, there is illustrated a detailed circuit diagram of RF amplifier 17 and mixer 20, respectively. Basically, amplifier 17 is comprised of a dual gate MESFET transistor 111 having a source coupled to a bias resistor and coupling capacitor 112, and a drain coupled to an L-C bandpass circuit 113. The gain of transistor 111 is varied about the levels indicated in FIG. 4a by an automatic gain control signal AGCRF. Signal AGCRF couples to a gate of transistor 111 through a voltage dividing network 114.

Similarly, mixer 20 is comprised of a single gate MESFET transistor 121 having a source coupled to a bias resistor and coupling capacitor 122, and a drain coupled to an L-C bandpass circuit 123. Circuit 123 has a center frequency of 330 MHZ. The mixing signals

from VHF VCO 22 are coupled to the gate of transistor 121 via an RC circuit 124.

It should be emphasized that the Schottky barrier gate structure of the MESFET transistors of amplifier 17 and mixer 20 yield significant performance improvements over other known devices. Narrow gate depletion mode MOSFET devices have a high frequency response, but they approximate square-law transfer characteristics over only a very narrow range of gate bias. Since departure from square-law operation results in cross modulation and intermodulation distortion, the devices are restricted to a small dynamic range. In comparison, the MESFET devices 111 and 121 have nearly ideal square-law characteristics over the entire operating range indicated in FIG. 4a. JFET devices also have a good square-law transfer characteristic, but their high frequency performance is greatly reduced because of parasitic capacitance and process difficulties that limit their usable geometries.

Referring now to FIG. 9a, there is illustrated a completed MESFET device suitable for use as mixer 20. The MESFET of FIG. 9a has a closed gate 131. Gate 131 is approximately 80 mils in length. The gate metal is approximately 0.3 mils wide, while the Schottky barrier gate in contact with the gate metal is approximately 0.15 mils wide. The source consists of five fingers 132-136 having lengths of 32.7 mils, 4.1 mils, 4.1 mils, 7.7 mils and 12.1 mils respectively. The width of these fingers is approximately 0.3 mils. The drain consists of fingers 137-140.

FIG. 9b is an enlarged cross-sectional view taken along the line A—A of FIG. 9a. The MESFET is constructed on a silicon substrate 140 having a P-type impurity. The resistivity of substrate 140 is approximately 50 ohm-centimeters. Each of the source electrodes 132-136 couples to an N+ doped region 142-146 respectively. These doped regions have a resistivity of approximately 0.005 ohm-centimeters. The doped regions extend beyond their corresponding electrodes by approximately 0.3 mils and are separated from the spaced apart gate electrode by approximately 0.15 mils. Similarly, the metal electrodes 137-140 which form the drain electrodes are coupled to underlying N+ doped regions 147-150, respectively. The resistivity and geometry of the drain doped regions are similar to that of the source doped regions.

FIG. 9c is a greatly enlarged cross sectional view of one source drain pair within a dual gate MESFET device suitable for use with amplifier 17. The entire device is constructed similar to that of FIG. 9a, with the modification that two gates are interleaved between the source and drain electrodes. In FIG. 9c, source electrode 132a and drain electrode 137a correspond to electrodes 132 and 137 of FIG. 9a. Gate electrodes 131a and 131b occupy the space of electrode 131 in FIG. 9a.

FIG. 9d illustrates the I-V characteristics of the devices of FIG. 9a, while FIG. 9e illustrates their transconductance as a function of gate voltage. In an ideal squarelaw device, the drain current is proportional to the square of the gate voltage. And since transconductance equals the partial derivative of drain current with respect to gate voltage, the transconductance is directly proportional to gate voltage for ideal square-law operation. FIG. 9e demonstrates such a linear relation between the transconductance and gate voltage for the MESFET device of FIG. 9a.

Referring now to FIG. 10, a circuit diagram of switch 25 is illustrated. Switch 25 is constructed identical to the previously described switch 15 of FIG. 6. The signals on leads 24 and 26 are selectively coupled to the output lead 27 via DC control signals select VHF (SELVHF) and select UHF (SELUHF) respectively.

FIGS. 11a-11h illustrate the details of SWD filter 28. Filter 28 is comprised of a surface wave device chip 160 having an L-C input circuit 170, and an L-C output circuit 190 for impedance matching and phase shifting signals to and from SWD device 160. Leads 172-174 and 192-194 couple circuits 170 and 190 to SWD 160 as illustrated in FIG. 11a.

Device 160 has a magnitude-frequency characteristic as illustrated in FIG. 11h. In particular, device 160 attenuates signals which are 1.5 MHZ above the picture carrier of the desired channel by at least 65 db. At that frequency, the sound carrier of the channel adjacent to the desired channel is present. It is important to greatly attenuate this sound carrier because it is translated into the video signals of the selected channel by synchronous detector 42. That is, the sound carrier at 46.5 MHZ is translated to 1.5 MHZ by detector 42. In the receiver of FIG. 1, just perceptible picture interference occurs when the sound carrier at 46.5 MHZ is passed through SWD filter 28 with a magnitude that is within 36 dB of the picture carrier at 45 MHZ. Thus, SWD filter 28 with its greater than 65 dB adjacent channel rejection enables the receiver of FIG. 1 to have good picture reception even though signals at the filter input in an adjacent channel are much larger than signals in the desired channel. For example, if the signal level of the desired channel is −55 DBM at the input of filter 28, then the 46.5 MHZ sound carrier at the input of filter 28 can be as high as −26 dBm and the receiver of FIG. 1 will meet the requirement of a −36 dB difference between the in band and out band signals at the output of filter 28. By comparison, prior art television receivers typically have perceptable picture interference when the adjacent channel sound carrier is −40 dBm.

FIGS. 11c and 11d illustrate the physical structure of one embodiment of SWD 160. Basically, device 160 is comprised of a piezoelectric substrate 161 which in a preferred embodiment is made of quartz. Quartz has a desirable feature in that the velocity of surface waves through quartz is practically independent of temperature. The dependence of velocity on temperature is of considerable importance since velocity effects the center frequency of the filter as described below.

Three electrically independent conductors 162-164 are disposed on substrate 161. Each conductor has corresponding finger electrodes 162a-164a which are disposed on substrate 161 in a comb like fashion. Fingers 162a-164a are equally spaced. The distance between two consecutive fingers on the same conductor is one wavelength of the center frequency of the filter. The velocity of a surface wave in quartz is approximately 3,300 meters per second, and the center frequency of device 160 is approximately 330 MHZ. Thus, the distance between two consecutive 162a fingers for example is approximately 10×10^{-6} meters.

Conductors 162-164 are coupled to the input circuitry 170 of FIG. 11a via leads 172-174 respectively. Input circuitry 170 generates voltages on electrodes 162-164 which are 120° out of phase with each other. This phase relationship generates a unidirectional surface wave on substrate 161. That is, waves in the forward direction add constructively, while waves in the reverse direction add destructively. As a result, device 160 has a low insertion loss for signals in the passband.

In particular, the loss through filter 160 is no more than 3.5 DB. U.S. Pat. No. 3,686,518 issued Aug. 22, 1972 to Hartmann et al and assigned to Texas Instruments Incorporated includes additional structural details of unidirectional surface wave filter 160.

FIG. 11d illustrates a method for shaping the impulse response of surface wave filter 160. The method therein illustrated is known as the finger withdrawal method. It involved removal of groups of fingers from selected portions of substrate 161. Basically, the amplitude of the impulse response in those regions from which the fingers are removed is reduced below the value which it would have if the fingers had not been removed. The method thus provides the capability to control the relative amplitude of the impulse response along the length of substrate 161. The desired impulse response is obtained by taking the inverse Fourier transfer of the frequency response of FIG. 11b; and then fingers 162a–164a are selectively removed in accordance with the desired impulse response. Further details of the finger withdrawal method are contained in U.S. Pat. No. 3,946,342 issued Mar. 23, 1976 to Hartmann and assigned to Texas Instruments Incorporated.

FIGS. 11e–11h illustrate a method for determining the value of inductors 175–177 and capacitor 178 comprising input circuitry 170. As was previously described, one of the functions of circuit 170 is to generate voltages on leads 172–174 which are shifted in phase from each other by 120°. FIG. 11e illustrates that this 120° phase shift can be achieved by a circuit 180 which produces a 60° phase lag between the voltage on lead 172 and the voltage on lead 173, in combination with a grounding of lead 174. This point is further illustrated by the phase diagram of FIG. 11f.

A circuit for producing a 60° phase lag is illustrated in FIG. 11g. The circuit consists of a pi shaped R-L-C network consisting of a capacitor 181, and inductor 182, a capacitor 183, and a resistor 184. Included in FIG. 11g are two equations relating the angle of phase lag between leads 172 and 173 in terms of components 181–184. In the case at hand, the angle of phase is 60°, and resistor 184 is the resistance between the electrodes 162 and 163 divided by 2. Thus, utilizing the equations of FIG. 11g, values for $|X_1|$ and $|X_c|$ can be calculated.

Impedance 183 is physically implemented by coupling inductor 175 across leads 173 and 174; and impedance 182 is physically implemented by coupling inductor 176 across leads 172 and 173 as illustrated in FIG. 11h. The parallel combination of inductor 175 and the capacitance between leads 173 and 174 due to the SWD electrodes is chosen to equal impedance 183. Similarly, the parallel combination of inductor 176 and the capacitance across leads 172 and 173 due to the SWD electrodes is chosen to equal impedance 182. Inductor 175 may typically equal 30 nanohenries, while inductor 176 may typically equal 35 nanohenries as an example.

Inductor 177 is then added between leads 172 and 174 while capacitor 178 is added between leads 172 and 29 so as to match the impedance between leads 172 and 174. Typically, inductor 177 is approximately 25 nanohenries, and capacitor 178 is 10 picofarads.

To this point, the discussion with reference to FIGS. 11a–11h has concentrated primarily on the structure of input circuitry 170 and the input transducer of SWD device 160. It will be understood however, that surface wave device 160 also has an output transducer on substrate 161 of a construction similar to that of the input transducer. Also, output circuitry 190 of FIG. 11a is constructed similar to input circuitry 170.

Referring next to FIGS. 12a and 12b, a circuit diagram of RF amplifier 29 is therein illustrated. Amplifier 29 is comprised of stages 200 and 201. Stage 200 includes a single bipolar transistor 202 as its amplifying element, whereas stage 201 includes several DC coupled bipolar transistors in an integrated circuit 203 as its amplifying element. Circuit 203 is indicated in FIG. 12a as a single circuit element, and is shown in detail in FIG. 12b.

Amplifier 29 provides a maximum gain of 30 DB to signals on its input lead 30. A small portion of the gain is provided by stage 200 which has a relatively good noise figure, whereas the remainder of the gain is provided by stage 201. AGC control signal AGCIF couples to stages 200 and 201 as indicated in FIG. 12a through bias circuits 216 and 217. The magnitude of signal AGCIF varies so as to keep the output of amplifier 29 at approximately −26.5 DBM. The variation of gain versus input signal strength for amplifier 29 was previously indicated in FIG. 4a.

The output of amplifier 29 couples to mixer 32 which is illustrated in the circuit diagram of FIG. 13. The basic mixing circuit element utilized therein is a bipolar transistor 220. Transistor 220 has an emitter 221 which is coupled to the output of amplifier 29. Similarly, transistor 220 has a base 223 which is coupled to SWD oscillator 35. Oscillator 35 generates mixing signals of 285 MHZ on the base 223 of transistor 220. As a result, sum and difference frequencies are generated on the collector 225 of transistor 220. Collector 225 is coupled to an LC tank circuit 226 having a resonant frequency of about 45 MHZ. The output of tank circuit 226 is coupled via a tapped transformer 227 to lead 37 and signals $S_4(f)$ are generated thereon.

Referring next to FIGS. 14a and 14b, a circuit diagram of oscillator 35 and a schematic diagram of acoustic surface wave resonator 36 are illustrated. Oscillator 35 includes a bipolar transistor 231 as the amplifying element. Transistor 231 has a grounded base 232, an emitter coupled via lead 233 to the input of resonator 36, and a collector coupled through a capacitor 234 to the output of resonator 36. An LC circuit 236 couples a DC supply voltage VDD to the collector of transistor 231 and also provides an output signal on lead 34.

Surface wave resonator 36 has a relatively high resonant frequency of 285 MHZ. Thus it is relatively small in size. Device size decreases as the resonant frequency increases. Typically, the SWD 36 of FIG. 14b is only approximately 0.10 inches in length. SWD 36 also has good long term frequency stability. This is because the resonator has a large Q. Typically, its Q is greater than 15,000. Q is the ratio of energy stored to energy dissipated per cycle within the device.

SWD resonator 36 is comprised of a piezoelectric substrate 240 which in the preferred embodiment is made of quartz. Reflective grating structures 241 and 242 are disposed at opposite ends of substrate 240. These grating structures form discontinuities in the surface of the substrate 240 which reflect surface waves thereon. Gratings 241 and 242 may be comprised of grooves, or alternatively bars of gold or copper for an example. The bars are spaced apart by one half wavelength of the resonant frequency. Typically, 250 to 400 bars are contained within each of the grating structures 241 and 242. The Q of the resonator increases as the number of bars increases. Also, as previously pointed

out, the velocity of the surface waves in quartz is relatively insensitive to temperature change. Thus, the resonant frequency or resonator 36 has low temperature drift. Typically, the resonant frequency varies less than 20 KHZ over a temperature range of 0-70° C.

An input transducer 243 and an output transducer 244 are disposed on substrate 240 in the space between grating structures 241 and 242. Lead 233 couples to input transducer 243, and lead 235 couples to output transducer 244. Transducers 243 and 244 are comprised of number of interleaved fingers which are placed at the peaks of the resonating standing wave that is set up by reflective grating structures 241 and 242. Typically, 60 fingers are on each transducer. Further details of resonator 36 are included in U.S. Pat. No. 3,886,504 issued to Hartmann et al. on May 27, 1975 and assigned to Texas Instruments Incorporated.

FIG. 15 is a circuit diagram of amplifier 38. Amplifier 38 is basically comprised of a circuit 250 which was previously illustrated in detail in FIG. 12*b*. Nodes A-G of circuit 250 as illustrated in FIG. 15 correspond to the node A-G as illustrated in FIG. 12*b*. Signals from mixer 32 couple to the input of circuit 250 through a capacitor 251. The gain of circuit 250 is automatically adjusted by a gain control signal AGCIF. Signal AGCIF is coupled to node C through an RLC circuit 252. The output of circuit 250 is coupled to lead 40 through an LC tank circuit 253.

FIGS. 16-19 are detailed circuit diagrams of the remaining portions of the channel selector of FIG. 1. The circuits utilize conventional components and are generally self-explanatory to those with ordinary skill in the art. Synchronous detector 42 is illustrated in FIG. 16. Detector 42 is essentially comprised of a commercially available chip MC1596. Chip MC1596 is described in the Linear Integrated Circuits Catalogue on pages 8-404 to 8-414 of Motorola's Semiconductor Data Library, Volume 6, Series, 1975. Details of the circuit are given on page 8-411 in FIG. 23 of the cited reference. Chip MC1596 is appropriately biased at each of its inputs and outputs by RLC circuits 261-269 as illustrated in FIG. 16. The biasing required by the component is also described in the cited reference.

Oscillator 45 is illustrated in FIG. 17. Oscillator 45 has a MESFET transistor 270 as the amplifying element, and frequency determining L-C feedback networks 271 and 272. Two separate output signals are provided by oscillator 45. One of the output signals is generated on lead 47 and the other signal is generated on lead 44. Leads 44 and 47 are separated by an RLC phase shifting network 273 which generates approximately a 90° phase difference between the two signals. Phase shifting network 273 insures that the picture carrier on lead 40 is in phase with the oscillator signal on lead 44. The phase detecting circuit 43, as illustrated in FIG. 18, generates phase detection signals PD1 and PD2 which lock the oscillator signal on lead 47 90° out of phase with the picture carrier on lead 40. Circuit 273 reinserts this 90° phase shift.

The phase detector 43 of FIG. 18 also utilizes chip MC1596 as was utilized in sync detector 42. Bias networks 281-287 are applied to chip MC1596 to achieve the phase detecting function. Ramp generator 54 and loop filter 49 couple to outputs of chip MC1596 as illustrated in FIG. 18. The combination generates phase detecting signals PD1 and PD2 on leads 51.

FIG. 19 illustrates a circuit diagram of VHF-VCO 22. VHF-VCO 22 utilizes phase signals PD1 and PD2 in combination with a third signal PD3 to generate the selectable frequency L0 signals on lead 23. Signal PD3 is a multilevel analog signal which provides a coarse voltage level of a unique value for each channel to be selected. Signal PD3 is generated by controller 52 in response to manually operated channel selection switches 56 as is previously described in conjunction with FIG. 1. Signals PD1-PD3 are coupled across a varactor diode 290 through summer 50. Diode 290 in combination with capacitors 291 and a micro strip 292 form the frequency determining circuit of VHF-VCO 22. Signal amplification within VCO 22 is provided by a bipolar transistor 293. Transistor 293 is selectively enabled or disabled by a control signal "enable VHF" (ENVHF). Signal ENVHF couples to the base of transistor 293 through a resistor 294 and to the collector of transistor 293 through an inductor 295. It is selectively energized in response to the position of the channel selection switches 56.

A complete television receiver which incorporates the channel selector of FIG. 1 is illustrated in block diagram form in FIG. 20. The channel selector has input leads 11 and 61 for receiving VHF and UHF television signals as previously described. The channel selector also has input leads 53 for receiving coarse analog voltages indicating the selected channel. The RF section of the channel selector frequency shifts the selected channel to approximately 330 MHZ, while the IF section of the channel selector filters the channel at 330 MHZ and frequency shifts the filtered channel to baseband.

Lead 46 is the audio output of the channel selector. As previously described, the sound carrier of the selected channel is generated on lead 46 at 4.5 MHZ. Lead 46 is coupled to the input of an audio demodulator 301. Demodulator 301 generates signals on a lead 302 with an amplitude that is proportional to the frequency of the frequency modulated signals on lead 46. This frequency demodulation process may be implemented by a variety of circuits that are well known in the art. The demodulated signals on lead 302 couple to the input of a speake. 303 where they are electromechanically converted to audible sounds.

Lead 72 is the composite video output of the channel selector. That is, signals on lead 72 include frame synchronizing information and video information of the selected channel. Lead 72 couples to the input of a video processing unit 304. Video processor 304 separates the picture signals from the frame synchronizing signals. The picture signals are generated on a lead 305 which couples to the electron gun input of a picture tube 306. The frame synchronizing signals are generated on a lead 307 which couples to the input of a drive circuit 308. Drive circuit 308 generates horizontal and vertical synchronizing signals on a lead 309 which couples to electron beam deflection circuitry 310 of picture tube 306. Drive circuit 308 also generates horizontal synchronizing signals on a lead 311 which couples to an input of picture tube high voltage generator 312. Additionally, drive circuit 308 generates synchronizing signals on lead 73 which couples to the AGC circuitry 74 of the channel selector.

Television receiver components 301-312 have been described in detail in many prior art publications. See for example the Fundamentals of Display System Design by Sol Sherr, 1970, published by Wiley-Interscience. A bibliography on pages 445-469 of the cited reference also includes many additional references.

FIG. 21 is a block diagram of a second embodiment of a channel selector constructed according to the invention. A significant portion of this second embodiment is similar in construction to that of the FIG. 1 embodiment. The similar portions are indicated by the identifying reference numerals.

One structural difference of the second embodiment is that it has only one VHF filter. That is, signals on VHF antenna 10 are coupled to a local oscillator mixer 320 through a fixed filter 321 which passes the entire VHF band of frequencies. A second difference is that the embodiment of FIG. 21 contains no RF amplifier. As a result, the system has improved intermodulation distortion and cross modulation distortion but has an increased noise figure. Still another difference is that the FIG. 21 embodiment contains only one RF MESFET mixer. A two by one switch 322 is provided having one input coupled to receive VHF signals on a lead 323 and a second input coupled to receive UHF signals on a lead 324. The output of switch 322 couples via a lead 325 to MESFET mixer 320.

The embodiment of FIG. 21 also includes a different means for generating the coarse channel selection voltages on leads 53. As FIG. 21 illustrates, controller 52 is comprised of a phase locked loop. The loop receives reference signals of a fixed frequency from a circuit 331, and simultaneously receives feedback signals from the LO VCO. The LO VCO signals are sent through a variable counter 332. Counter 332 divides by a number which is selectable via logic signals on leads 333. The signals on leads 333 are generated by a logic circuit 334 in response to logic signals received from the channel selection switches 56. A phase detector 335 compares the output signals of variable counter 332 to the reference signals, and generates phase detection signals for the LO VCO. Thus, a relatively high local oscillator frequency is generated when counter 332 is selected to divide by a relatively large number and vice versa.

Various embodiments of the invention have now been described in detail. However, many changes and modifications can be made to the above details without departing from the nature and spirit of the invention. For example, the IF frequency of the mixer output is not restricted to 330 MHZ. Other IF frequencies in the range of 300–400 MHZ may be employed. As another example, a bipolar RF amplifier may be substituted for the MESFET RF amplifier. This is because the mixer introduces cross modulation distortion and intermodulation distortion into the receiver to a much smaller degree than does the linear RF amplifier. Thus, utilization of a MESFET mixer yields a receiver having greatly reduced third order distortion even though the RF amplifier is bipolar. As another example, the surface wave device resonator utilized to generate 285 MHZ mixing signals may be comprised of a single transducer as opposed to a dual transducer. In a single transducer resonator, lead 233 couples to one set of electrodes on the transducer while lead 235 couples to the other set of electrodes. The single transducer is configured similar to transducer 243. As still another example, the channel selector of FIG. 1 or FIG. 21 may be readily adapted for use in systems other than television receivers. The channel selector has application wherever one channel of frequencies is desired to be selected from a plurality of nonoverlapping frequency channels. The information contained in the channels need not be television signals. Therefore, it will be understood that many changes and modifications can be made in the above details without departing from the nature and spirit of the invention. It is understood that the invention is not limited to said details except as set forth in the appended claims.

What is claimed is:

1. In a channel selector for a television receiver including a radio frequency section and an intermediate frequency section, the combination comprising:

spectrum filter means disposed in the radio frequency section for receiving radio frequency signals representative of a plurality of television channels and for filtering at least one frequency spectrum of television channels from the received signals;

at least one mixing means connected to the output of said spectrum filter means, said mixing means including a metal semiconductor field effect transistor and having first and second inputs coupled to simultaneously receive said frequency spectrum of television channels from said spectrum filter means and mixing signals of a selected intermediate frequency for frequency shifting the selected frequency channel of said frequency spectrum of television channels to a high intermediate frequency substantially greater than 45 MHZ;

channel selecting filter means having an input coupled to the output of said mixing means for filtering said selected frequency channel at said high intermediate frequency so as to pass said selected frequency channel;

second mixing means disposed in the intermediate frequency section and having first and second inputs coupled to simultaneously receive the filtered selected frequency channel at said high intermediate frequency from the output of said channel selecting filter means and mixing signals of a predetermined fixed frequency for frequency shifting said filtered selected frequency channel at said high intermediate frequency to a substantially lower second intermediate frequency; and

amplifier means having a relatively high gain disposed in said intermediate frequency section and connected to the output of said second mixing means for providing gain to said filtered selected frequency channel at the lower second intermediate frequency, the added gain provided by said amplifier means in said intermediate frequency section being substantially in excess of the total gain provided to the signal in the radio frequency section through said channel selecting filter means.

2. In a channel selector as set forth in claim 1, further including amplifier means having a relatively low gain disposed in said radio frequency section so as to be interposed between said spectrum filter means and said at least one mixing means, said amplifier means in said radio frequency section having an input coupled to said spectrum filter means for receiving said frequency spectrum of television channels from said spectrum filter means and an output coupled to the first input of said at least one mixing means for transmitting low gain amplified signals of said frequency spectrum of television channels to said at least one mixing means.

3. In a channel selector as set forth in claim 2, wherein the total gain provided to the signal in the radio frequency section by said amplifier means disposed therein is less than 10dB.

4. In a channel selector as set forth in claim 1, wherein said spectrum filter means comprises at least one fixed filter having at least one predetermined frequency passband for receiving radio frequency signals representa-

tive of a plurality of television channels and for filtering at least one fixed frequency spectrum of television channels from the received signals.

5. In a channel selector as set forth in claim 1, wherein said channel selecting filter means comprises acoustic surface wave filter means.

6. In a channel selector as set forth in claim 1, wherein the high intermediate frequency produced from said at least one mixing means is between 300 MHZ and 400 MHZ.

7. In a channel selector as set forth in claim 6, wherein the lower second intermediate frequency produced

from said second mixing means is approximately 45 MHZ.

8. In a channel selector as set forth in claim 1, further including additional amplifier means connected between said channel selecting filter means and the first input of said second mixing means for providing gain to said filtered selected frequency channel passed by said channel selecting filter means prior to the introduction of said filtered selected frequency channel at the high intermediate frequency to the first input of said second mixing means.

* * * * *

TAB 12

United States Patent [19] | [11] 4,340,975
Onishi et al. | [45] Jul. 20, 1982

[54] MICROWAVE MIXING CIRCUIT AND A VHF-UHF TUNER HAVING THE MIXING CIRCUIT

[75] Inventors: Hiroshi Onishi, Kawasaki; Sadahiko Yamashita, Sagamihara, both of Japan

[73] Assignee: Matsushita Electric Industrial Company, Limited, Osaka, Japan

[21] Appl. No.: 195,314

[22] Filed: Oct. 8, 1980

[30] Foreign Application Priority Data

Oct. 9, 1979 [JP] Japan 54-130573
Apr. 14, 1980 [JP] Japan 55-49377

[51] Int. Cl.³ H03D 7/02; H03D 7/14; H04B 1/26; H04N 5/44

[52] U.S. Cl. 455/315; 358/191.1; 455/327; 455/330; 455/331

[58] Field of Search 455/315, 316, 318, 319, 455/320, 323, 325, 326, 327, 330, 331; 358/191.1

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Primary Examiner—Howard Britton
Attorney, Agent, or Firm—Lowe, King, Price & Becker

[57] ABSTRACT

In a double superheterodyne tuner for receiving TV channel signals, a frequency in a range defined between 2520 and 2700 MHz is selected as the first intermediate frequency, while the first intermediate frequency is obtained by subtracting a desired channel frequency from a first local oscillator frequency. The tuner comprises first and second mixing circuits respectively used for effecting frequency conversion, where each of these mixing circuits has parallel coupled lines so that each mixing circuit operates as a balance-to-unbalance transformer. The second mixing circuit further comprises a short-circuit line placed between the parallel coupled lines. The tuner also comprises a band pass filter having a band width less than 5 MHz, which selectively transmits the first intermediate frequency signal from the first mixing circuit.

29 Claims, 21 Drawing Figures

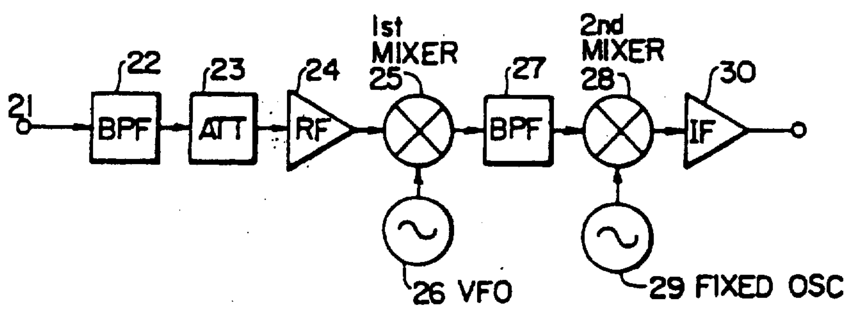


FIG. 1 PRIOR ART

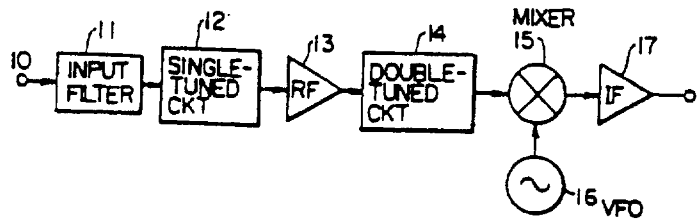


FIG. 2

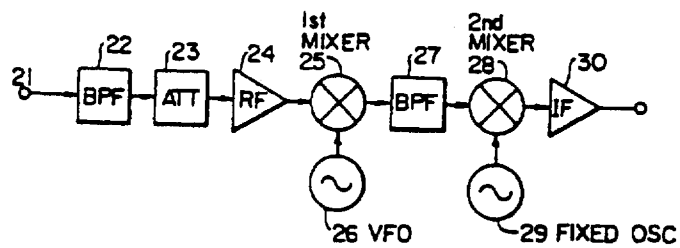


FIG. 3

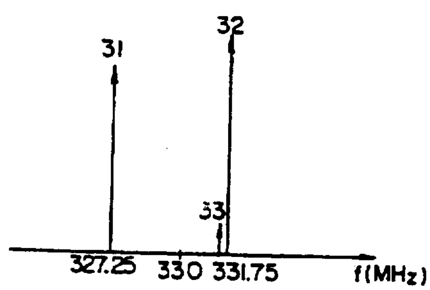


FIG. 4 PRIOR ART

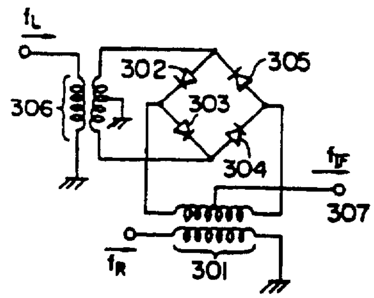


FIG. 5 PRIOR ART

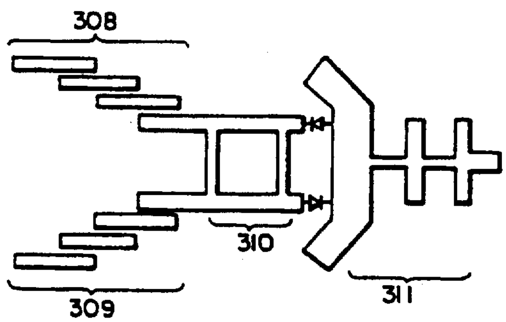
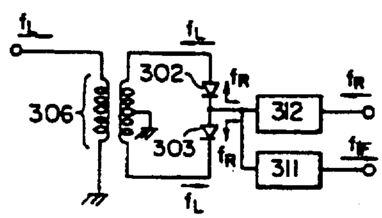
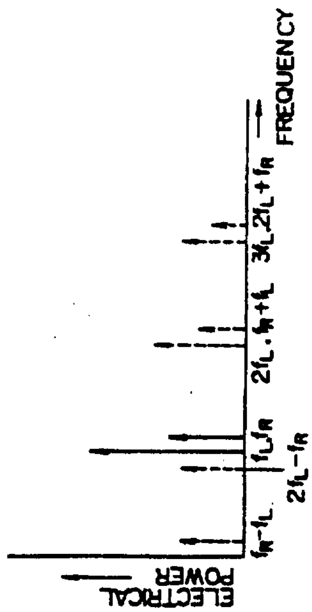


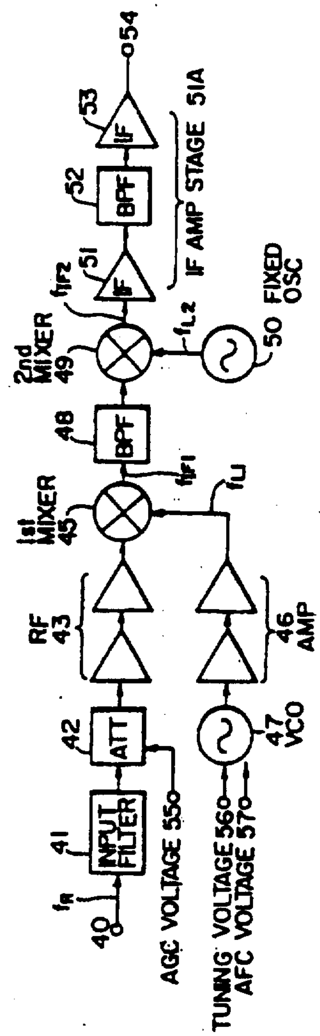
FIG. 6 PRIOR ART





ELECTRICAL POWER
FREQUENCY

FIG. 7



INPUT FILTER
ATT
RF
AMP
VCO
1st MIXER
BPF
2nd MIXER
IF
FIXED OSC
IF AMP STAGE
AGC VOLTAGE
TUNING VOLTAGE
AFC VOLTAGE

FIG. 8

FIG. 9

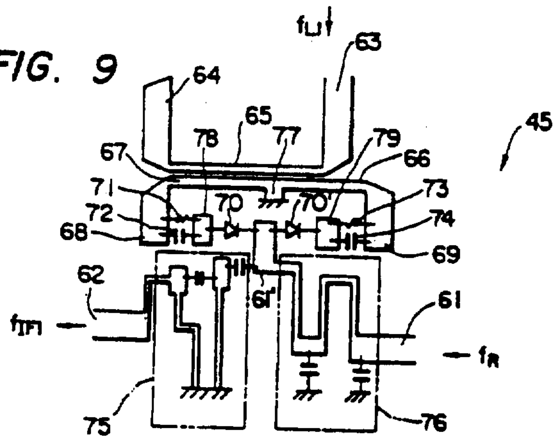


FIG. 10

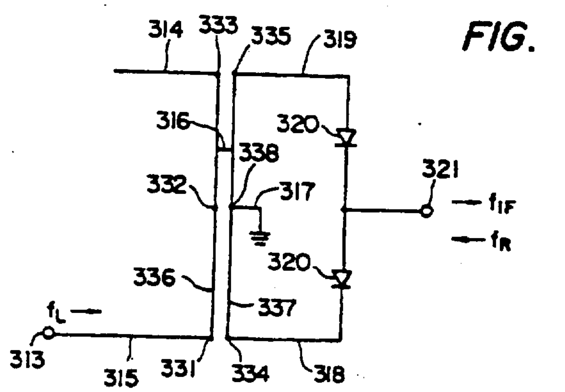


FIG. 11A

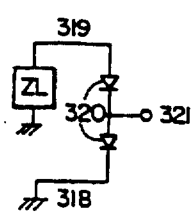


FIG. 11B

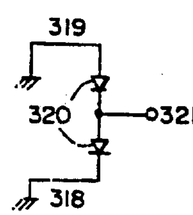
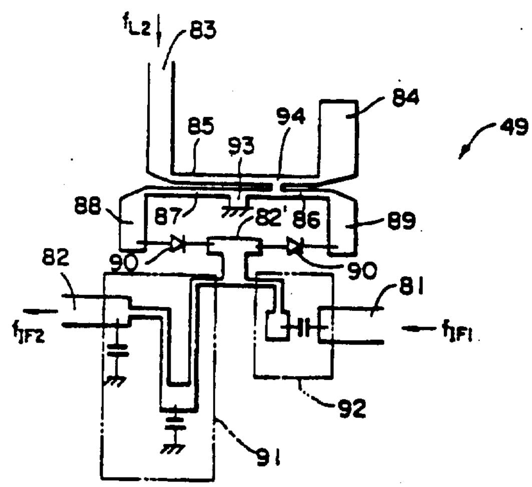
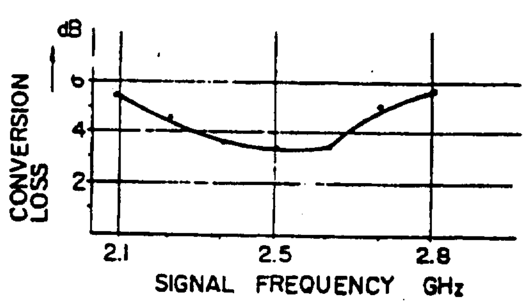


FIG. 12



f_{L2}
83
84
85
94
93
49
88
87
82'
86
89
82
90
90
81
f_{IF2}
f_{IF1}
92
91

FIG. 13



dB
CONVERSION LOSS
6
4
2
2.1
2.5
2.8
SIGNAL FREQUENCY GHz

FIG. 14A

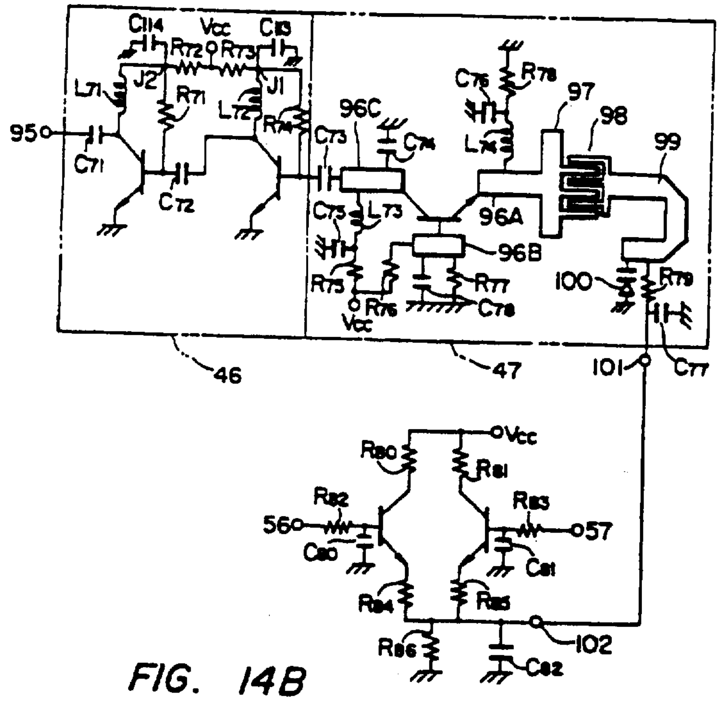


FIG. 14B

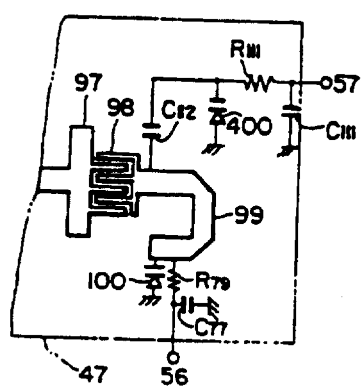


FIG. 15

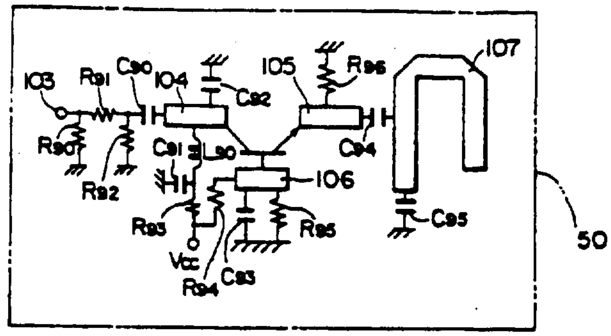

103
R91
C90
104
C92
105
R96
107
R90
R92
C91
L90
C94
106
R93
Vcc
R94
C93
R95
C95
50

FIG. 16

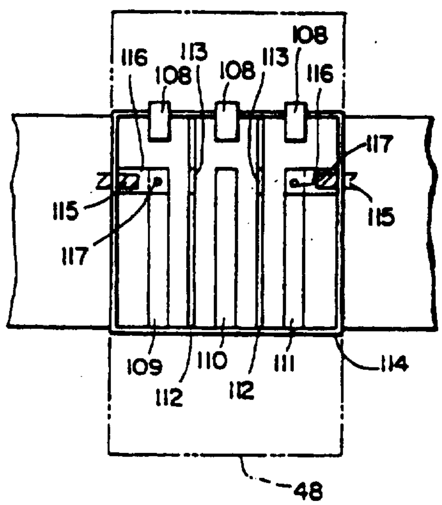

108
116
113
108
108
113
116
117
115
117
115
109
110
111
112
112
114
48

FIG. 17

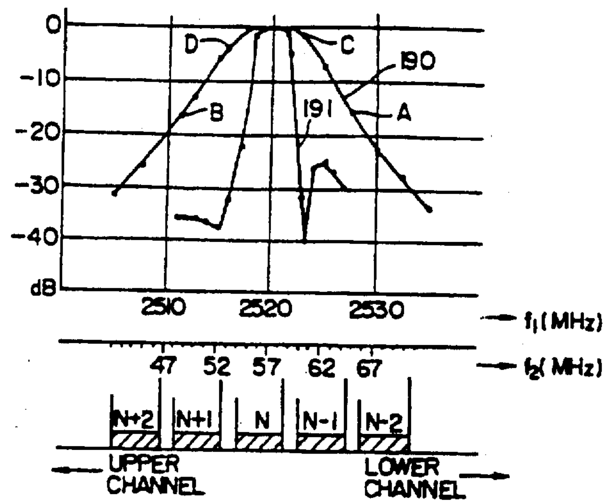

0
-10
-20
-30
-40
dB
D
C
190
B
191
A
2510
2520
2530
f1 (MHz)
47 52 57 62 67
f2 (MHz)
N+2
N+1
N
N-1
N-2
UPPER CHANNEL
LOWER CHANNEL

FIG. 18

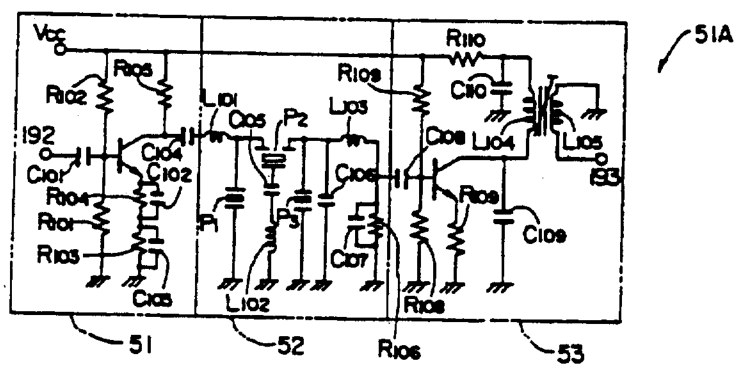
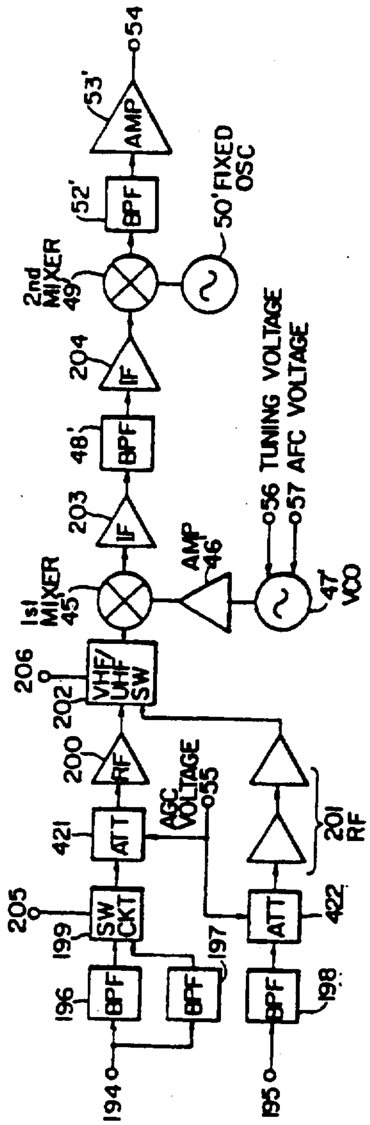

Vcc
R110
R105
R109
R102
L101
C105
P2
L103
C110
192
C104
C108
L104
L105
C101
C102
C106
193
R104
R108
R101
P1
P3
C109
R103
C107
C103
L102
R106
51
52
53
51A

FIG. 19


194
195
BPF
196
BPF
197
BPF
198
SW CKT
199
205
ATT
421
ATT
422
AGC VOLTAGE
55
RF
200
201 RF
VHF/ UHF SW
202
206
1st MIXER 45
AMP 46
VCO 47
56 TUNING VOLTAGE
57 AFC VOLTAGE
IF
203
BPF
48'
IF
204
2nd MIXER 49
50' FIXED OSC
BPF
52'
AMP
53'
54

MICROWAVE MIXING CIRCUIT AND A VHF-UHF TUNER HAVING THE MIXING CIRCUIT

FIELD OF THE INVENTION

The present invention relates generally to a tuner for selectively receiving a desired channel signal among a plurality of received TV signals, and particularly, the present invention relates to VHF-UHF all-channel tuner of double superheterodyne type.

BACKGROUND OF THE INVENTION

In order to receive both VHF and UHF TV broadcast signals it has been required to utilize two independent tuners hitherto, where these two tuners have to be switched from one to another. However, recently, an all-channel electronic tuner of double superheterodyne type has been proposed so that all channel signals can be simply selected without switching from one tuner to another. As will be described with reference to the accompanying drawings, such an electronic all-channel tuner has a disadvantage that spurious signals would occur due to harmonics of a local oscillator output signal.

In a tuner of the superheterodyne or double superheterodyne type, mixer or mixers, which convert the frequency of an incoming signal into a desired intermediate frequency, play an important role where the characteristics of mixer(s) have to meet various requirements. Although there are many known compact mixers, in these conventional mixers, image frequency components as well as the harmonic components of the local oscillator frequency are not positively processed or treated. As a result, conventional mixers have to suffer conversion loss of 6 to 10 dB.

SUMMARY OF THE INVENTION

The present invention has been developed in order to remove the above-mentioned disadvantages and drawbacks inherent in the conventional tuners and mixers.

It is, therefore, a primary object of the present invention to provide a new and useful VHF-UHF all-band tuner for receiving TV broadcast signals, in which spurious signal interference caused by channel signals other than a desired channel signal is suppressed to the utmost.

Another object of the present invention is to provide a VHF-UHF all-band tuner for receiving TV broadcast signals, having low-noise and low-distortion characteristics.

A further object of the present invention is to provide a mixing circuit for a VHF-UHF all-band tuner, in which the conversion loss is remarkably improved.

A still further object of the present invention is to provide a single-balanced mixing circuit for a VHF-UHF all-band tuner, in which the sum components of the local oscillator frequency and the incoming signal frequency, as well as the second harmonic component of the local oscillator frequency is suppressed by means of an original balance-to-unbalance transformer formed in a microwave integrated circuit.

BRIEF DESCRIPTION OF THE DRAWINGS

These and other objects and features of the present invention will be more readily apparent from the following detailed description of the preferred embodiments taken in conjunction with the accompanying drawings in which:

FIG. 1 is a block diagram of a conventional VHF electronic tuner;

FIG. 2 is a schematic block diagram showing the fundamental structure of a double superheterodyne all-band tuner;

FIG. 3 is a chart showing the relationship between a first intermediate frequency of 330 MHz and interference signal caused by channel signals other than a desired channel signal;

FIG. 4 is a circuit diagram of a conventional double-balanced mixer with a balance-to-unbalance transformer;

FIG. 5 is a top plan view of a conventional single-balanced mixer, which is in the form of a microwave integrated circuit, having a directional coupler,

FIG. 6 is a circuit diagram of a conventional single-balanced mixer with a balance-to-unbalance transformer;

FIG. 7 is a spectrum chart showing frequencies which occur in a mixing diode;

FIG. 8 is a schematic block diagram of a first embodiment of the VHF-UHF all-band tuner according to the present invention;

FIG. 9 is a schematic diagram of the first mixing circuit shown in FIG. 8;

FIG. 10 is a schematic diagram showing the fundamental structure of the second mixing circuit shown in FIG. 8;

FIG. 11A is an equivalent circuit of the balance-to-unbalance transformer of FIG. 10 in the absence of the short-circuit stripline at the sum frequency of the local oscillator frequency and an input signal frequency, and at the second harmonic frequency of the local oscillator frequency;

FIG. 11B is an equivalent circuit of the balance-to-unbalance transformer of FIG. 10 in the presence of the short-circuit stripline at the same frequencies.

FIG. 12 is a schematic diagram of the second mixing circuit shown in FIG. 8;

FIG. 13 is a graphical representation showing the conversion loss of the second mixing circuit shown in FIG. 12;

FIG. 14A is a schematic circuit diagram of the combination of the voltage controlled oscillator equipped with an AFC circuit; and the local oscillator frequency amplifier connected thereto shown in FIG. 8;

FIG. 14B is a schematic circuit diagram of a modified AFC circuit which may be used in combination with the voltage controlled oscillator of FIG. 14A;

FIG. 15 is a schematic circuit diagram of the fixed frequency oscillator shown in FIG. 8;

FIG. 16 is a schematic top plane view of the fixed frequency band pass filter shown in FIG. 8;

FIG. 17 is a graphical representation showing the relationship between the frequency response of the frequency fixed band pass filter of FIG. 16 and the frequency response of the second intermediate frequency amplifying stage of FIG. 8;

FIG. 18 is a schematic circuit diagram of the second intermediate frequency amplifying stage shown in FIG. 8; and

FIG. 19 is a schematic block diagram of a second embodiment of the VHF-UHF all-band tuner according to the present invention.

TUNE 037132

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

Prior to describing the preferred embodiments of the present invention, conventional tuners and mixing circuits used in tuners will be discussed for a better understanding of the objects of the present invention.

FIG. 1 illustrates a block diagram of a VHF electronic tuner which has been used up to this time. An incoming or received signal is applied to an input filter 11 from a input terminal 10 so that interference signal in the intermediate frequency range, FM signals and the like which may interfere the TV intermediate frequency signal are eliminated from the incoming signal to be applied to a single-tuned circuit 12. In the single-tuned circuit 12, a desired channel is selected by means of a variable resonance element constructed of a varactor (not shown), and then the selected channel signal is applied to a radio frequency (RF) amplifier 13 to be amplified. The amplified signal is applied to a double-tuned circuit 14, in which two varactors (not shown) are used as variable resonance elements so that undesired signals are eliminated. The output signal of the double-tuned circuit 14 is then applied to an input terminal of a mixing circuit 15 which is also responsive to an oscillating signal from a variable frequency oscillator 16 having a varactor (not shown). Thus, the output signal of the double-tuned circuit 14 is converted to an intermediate frequency (IF) signal in the mixing circuit 15, and this IF signal is applied to an intermediate frequency (IF) amplifier 17 to be amplified therein.

In the above-described conventional tuning circuit, one or more varactors are required in each of the above-mentioned single-tuned circuit 12, double-tuned circuit 14 and the variable frequency oscillator 16, and tracking adjustment between the respective circuits is necessary. Furthermore, in order to receive both VHF and UHF signals two independent tuning circuits are required. In addition, it is necessary to change the tuning frequency and oscillating frequency by means of a switching diode to receive both a lower frequency band and a higher frequency band in VHF range.

Recently, an all-channel tuner of the double superheterodyne type has been proposed in order to remove the above-mentioned disadvantages and drawbacks inherent to the conventional electronic VHF-UHF tuner. There are two types in the already proposed double super-heterodyne tuners as follows: The first type is that the first intermediate frequency of f_{IF1} equals the difference between the local oscillator frequency f_{L1} and the incoming or received signal frequency f_R, namely $f_{L1}-f_R=f_{IF1}$, where f_{IF1} is selected between 300 and 400 MHz. The second type is that the first intermediate frequency f_{IF1} equals the sum of the local oscillator frequency f_{L1} and the incoming signal frequency f_R, namely, $f_{L1}+f_R=f_{IF1}$, where f_{IF1} is 3000 MHz. Both of these types have drawbacks as will be described in detail with reference to FIG. 2 and FIG. 3 hereinbelow.

FIG. 2 shows a basic structure of an all-band tuner of the double superheterodyne type. In FIG. 2, a reference numeral 21 designates an input terminal connected to a fixed band pass filter 22 having a necessary range, such as VHF and UHF. A reference numeral 23 designates a variable attenuator which follows the band pass filter 22, and the output terminal of the variable attenuator 23 is connected to an input terminal of an RF amplifier 24 the output terminal of which is connected to a first input

terminal of a first mixing circuit 25. With this arrangement, a high frequency incoming or received signal applied to the input terminal 21 is filtered, attenuated, and then amplified to be applied to the first mixing circuit 25. The first mixing circuit 25 receives via its second input terminal a local oscillator signal from a variable frequency oscillator 26 so that the local oscillator frequency is mixed with the incoming high frequency applied via the first input terminal of the mixing circuit 25. As a result, the incoming high frequency signal corresponding to a desired TV channel is converted into a first intermediate frequency. The output terminal of the first mixing circuit 25 is connected to an input terminal of a second band pass filter whose filtering frequency is fixed, so that signals in the vicinity of the first intermediate frequency are selectively transmitted to the following circuits to reduce interference due to signals of other channels. The output terminal of the second band pass filter 27 is connected to a first input terminal of a second mixing circuit 28 which also receives via its second input terminal a local oscillator signal from a fixed oscillator 29. The first intermediate frequency applied to the first input terminal of the second mixing circuit 28 is, therefore, converted into a second intermediate frequency which corresponds to the intermediate frequency of a TV set. An intermediate frequency amplifier 30 follows the second mixing circuit 28 for amplifying the second intermediate frequency.

In the above-described basic structure of the tuner, characteristics of respective circuits are very important when determining the total characteristic of the tuner, while the interference-rejection characteristics of the tuner is greatly affected by the value of the first intermediate frequency.

Suppose the first intermediate frequency is set to a frequency, for instance, of 330 MHz, in a range between 300 and 400 MHz, and the aforementioned first mixing method or type is adopted. Namely, the first intermediate frequency f_{IF1} is obtained by detecting the difference between the local oscillator frequency f_{L1} from the variable oscillator 26 and the high frequency incoming signal frequency f_R.

Considering signals of channels other than a desired channel as interference signals, the frequency range of the interference signals is, for instance, in case of TV bands in the United States of America, is between 54 to 890 MHz. The frequency or frequencies of the interference signal or signals will be expressed in terms of f_u hereinbelow throughout the specification.

High harmonic interferences, such as $f_{L1}-2f_u$, $f_{L1}-3f_u$ and the like to be converted into f_{IF1} when receiving a desired channel signal, necessarily occur regardless of which frequency is selected as the first intermediate frequency. However, such interferences may be removed by adding fixed band pass filters respectively for the VHF higher band, VHF lower band, and UHF band to the input side.

If the first intermediate frequency and the mixing method are respectively selected as described in the above, a frequency expressed in terms of $2f_u-f_{L1}$ becomes an interference signal in the UHF band. The degree of the interference due to this interference signal is great because the interference corresponds to the difference between the second harmonic of the interference signal and the fundamental wave of the local oscillator frequency signal. This point will be further de-

scribed taking the case of receiving TV signals in the United States of America.

Suppose channel 56 whose center frequency is 725 MHz is the desired one. Since the first intermediate frequency is 330 MHz, the local oscillator frequency f_{L1} applied to the first mixing circuit 25 is 1055 MHz. If there exist signals of channel 51, whose video carrier frequency is 693.25 MHz and sound carrier frequency is 697.75 MHz, because of mixing between the second harmonic of the channel 51 signals and the local oscillator signal, the second harmonic component of the video carrier signal of the channel 51 gets in the first intermediate frequency band of the desired channel. FIG. 3 illustrates the relationship between these signals. In FIG. 3, a reference numeral 31 designates a frequency-converted sound carrier signal of 327.25 MHz of the desired channel, and 32 is a frequency-converted video carrier signal of 331.75 MHz of the same desired channel. A reference numeral 33 designates an interference signal of 331.5 MHz due to the second harmonic of the video carrier signal of the channel 51.

Furthermore, when the first intermediate frequency (330 MHz) and the mixing method are respectively set as mentioned in the above, when receiving signals of channel 46 whose center frequency is 665 MHz, video carrier frequency is 663.25 MHz, and sound carrier frequency is 667.75 MHz, the second harmonic of the video carrier frequency of the desired channel per se is frequency converted so that interference would occur having the relationship in frequency as illustrated in FIG. 3.

The above-described interference inevitably occurs because TV channel signals of UHF band are frequency converted into the first intermediate frequency between 300 and 400 MHz, and the interference will result in beat interference in a TV picture.

Next, let us suppose that the frequency of the output signal of the first mixing circuit 25, namely, the first intermediate frequency, is, for instance, 3000 MHz, where the first intermediate frequency f_{IF1} equals the sum of the local oscillator frequency f_{L1} and the incoming signal frequency f_R.

When mixing is effected in the above-mentioned manner, the difference between the second and third harmonics of the interference signal frequency f_u and the second harmonic of the local oscillator frequency, i.e. $2f_{L1}-2f_u$ and $2f_{L1}-3f_u$, get in the first intermediate frequency band of the desired channel. This point will be further described taking the case of receiving TV channel signals in the United States of America.

Assuming that the tuner has been tuned to channel 50 the center frequency of which is 689 MHz, the local oscillator frequency applied to the first mixing circuit 25 is 2311 MHz. If there exist signals of channel 70 whose video carrier frequency is 807.25 MHz, and sound carrier frequency is 811.75 MHz, the difference between the second harmonic of the local oscillator frequency and the sound carrier frequency of the channel 70 is 2998.5 MHz so that this gets in the first intermediate frequency band of the channel 50, which is to be received, resulting in interference.

From the above, it will be understood that the present invention aims to remove the above described disadvantages inherent to the conventional TV tuner, and the invention provides a VHF-UHF all-band tuner of double superheterodyne type suitable for a TV set. One of the features of the present invention is that various interferences which could not be avoided according to the prior art, can be eliminated by properly selecting the first intermediate frequency and also by adopting a suitable mixing technique. Another feature of the present invention is that the frequency of the output signal of the first mixing circuit 25, i.e. the first intermediate frequency, is set between 2520 and 2700 MHz, and the first intermediate frequency is obtained by detecting the difference between the local oscillator frequency, which is higher than the first intermediate frequency, and the incoming or received signal frequency.

The manner by which the above-mentioned various interferences are eliminated will be described in detail hereinbelow. Considering signals of TV channels other than a desired channel as interference signals, the ranges of the interference signals, for instance in connection with TV bands in the United States of America, are respectively defined between 54 and 85 MHz of the VHF lower band, between 174 and 216 MHz of the VHF higher band, and between 470 and 890 MHz of the UHF band. In the same manner, in the case of TV bands in Japan, the interference signal ranges are respectively defined between 90 and 108 MHz of the VHF lower band, between 170 and 222 MHz of the VHF higher band, and between 470 and 770 MHz of the UHF band.

Spurious interferences which occur in the first mixing circuit 25 are caused, except for beat interferences between channels, by the fact that spurious signals produced by mixing the local oscillator frequency f_{L1} applied to the first mixing circuit 25 with the harmonics of the interference signal, get in the first intermediate frequency band of the desired channel signal. This relationship will be given by the following formula:

$$|mf_{L1} \pm nf_u| = f_{IF1}$$

wherein "m" and "n" are respectively 0, 1, 2 or 3 . . . .

In the above, the harmonics of f_{L1} and f_u are respectively considered to be as far as the third order. Namely, harmonics of more than the third order do not have to be considered since the presence of such higher harmonics means that the system is saturated.

The above-mentioned various interference components will be classified in accordance with the order of the harmonics of the local oscillator frequency f_{L1} as follows:

(1) Interferences caused by the fact that the second and third harmonics per se of the interference signal get in the first intermediate frequency band:

$$2f_u \text{ and } 3f_u$$

(2) Interferences caused by mixing the fundamental wave of the local oscillator frequency with the second and third harmonics of the interference signal:

$$|f_{L1} \pm 2f_u| \text{ and } |f_{L1} \pm 3f_u|$$

(3) Interference caused by mixing the second harmonic of the local oscillator frequency with the fundamental wave, second harmonic and the third harmonic of the interference signal:

$$|2f_{L1} \pm f_u|,\ |2f_{L1} \pm 2f_u| \text{ and } |2f_{L1} \pm 3f_u|$$

(4) Interference caused by mixing the third harmonic of the local oscillator frequency with the fundamental wave, second harmonic and the third harmonic of the interference signal:

$|3f_{L1} \pm f_u|$, $|3f_{L1} \pm 2f_u|$ and $|3f_{L1} \pm 3f_u|$

Let us suppose that the first intermediate frequency is selected as 2700 MHz. Considering the case of receiving TV channel signals in the United States of America, the range of the local oscillator frequencies applied to the first mixing circuit 25 is from 2757 to 3587 MHz. Let us further assume that interference signals reside in the VHF lower band, VHF higher band, and the UHF band between the entire range defined between 54 and 890 MHz.

Since the interference signals reside in the above-mentioned frequency ranges, the above-mentioned $2f_u$ and $3f_u$ of the item (1) do not directly get in the first intermediate frequency of the desired channel signal at all. In the interferences caused by mixing the fundamental wave of the local oscillator frequency with the fundamental wave and the harmonics of the interference signal as described in the item (2), only $f_{L1}-2f_u$ and $f_{L1}-3f_u$ will be interference signals because of the frequency relationship between the signals to be mixed.

The spurious interferences due to harmonics cannot be avoided no matter what frequency is selected as the first intermediate frequency, and no matter what kind of mixing method or technique is adopted so that it is necessary to suppress the interference signals, which result in the spurious interferences due to harmonics, at the input side. The above-described interferences of items (3) and (4), namely, signals which get in the first intermediate frequency band of the desired channel by mixing the second or third harmonics of the local oscillator frequency applied to the first mixing circuit 25, do not occur because of the frequency relationship such that the second harmonic of the local oscillator frequency is between 5514 and 7174 MHz, the third harmonic of the same is between 8271 and 10761 MHz, and the interference signal is between 54 and 890 MHz.

In the same manner, when it is intended to receive TV channel signals in Japan with the first intermediate frequency of 2700 MHz, the local oscillator frequency applied to the first mixing circuit 25 is between 2793 and 3467 MHz so that interferences by the channel signals other than the desired channel are the same as in the case of receiving U.S. TV channel signals except for the beat interference between channels.

Since the TV broadcasting bands in Japan reside in a range of 90 to 770 MHz, if the first intermediate frequency is set to 2520 MHz, interferences by the channel signals other than the desired channel during receiving Japanese TV signals are the same as described in the above. In other words, the only problem is the interference caused by mixing the fundamental frequency of the local oscillator signal applied to the first mixing circuit 25 with the harmonics of the other channel signals.

Another feature of the present invention is that both the combination of the first mixing circuit of 2500 to 3500 MHz band and the variable frequency local oscillator for processing the first intermediate frequency, and the combination of the second mixing circuit and the fixed local oscillator for converting the first intermediate frequency into the intermediate frequency of a TV set, are integrally formed in microwave integrated circuits respectively. Because the main portions are formed in an integrated circuit, uniform characteristics throughout a number of products are obtained so that mass production is readily effected.

Furthermore, in accordance with the present invention a VHF-UHF all-band tuner having a low-noise characteristic and a low modulation distortion characteristic can be provided.

In accordance with a preferred embodiment, the noise figure in a VHF band is approximately 5 dB and in UHF band, the same is approximately 7 dB, while the input signal level of immediately adjacent or second adjacent channel signals, which give a cross modulation distortion of 1 percent to the desired channel signal is from −20 to −25 dBm. The low cross modulation distortion characteristic has been achieved by a radio frequency amplifier having a low distortion characteristic, and a diode single balanced mixer having an original structure.

The input signal level of the first mixer, by which a cross modulation distortion of 1 percent is given to the desired channels signal is approximately 0 dBm, and the detailed structure of the first mixer will be described later.

In addition, by properly selecting the band of the fixed band pass filter, which selectively transmits the first intermediate frequency, and by the provision of the second intermediate frequency having an original structure, the degree of the cross modulation distortion given by the immediately adjacent channel signals to the desired channel signal can be reduced to the same level as that by the second adjacent channel signals.

In the above, it has been described how the various interferences are eliminated. Now let us discuss some conventional mixing circuits for further making the objects of the present invention clear.

FIG. 4 is a circuit diagram of a conventional double-balanced mixer, also called a ring modulator, which has been widely used hitherto. In FIG. 4, reference numerals 301 and 306 designate balance-to-unbalance transformers, and in detail, the transformer 301 is used for receiving an incoming signal, while the other transformer 306 is used for receiving a local oscillator signal. Four diodes 302, 303, 304 and 305 constitute a bridge circuit between the secondary windings of the transformers 301 and 306. References f_R and f_L respectively indicate the incoming signal frequency and the local oscillator frequency. The secondary winding of the transformer 306 has a center tap connected to ground, while the secondary winding of the other transformer 301 has a center tap connected to an output terminal 307 from which an intermediate frequency signal will be derived.

FIG. 5 is a top plan view of a widely used conventional single-balanced mixer formed in a microwave integrated circuit (MIC), which mixer has a directional coupler of 3 dB. In FIG. 5, reference numerals 308 and 309 respectively designate band pass filters for the local oscillator signal and for the incoming signal. Each of the band pass filters 308 and 309 is connected to a directional coupler 310 of 3 dB which is connected via diodes (no numeral) to an intermediate frequency band pass filter 311. In this mixer circuit, since both branches of the directional coupler 310 require electrical lengths respectively corresponding to a quarter wavelength of the incoming signal frequency and a quarter wavelength of the local oscillator frequency, the mixer circuit tends to be bulky if the frequency is below several GHz.

FIG. 6 is a circuit diagram of a widely used conventional single-balanced mixer with a balance-to-unbalance transformer. The local oscillator frequency f_L is converted by means of a transformer 306 into balanced

outputs, and the balanced outputs are respectively applied to mixing diodes 302 and 303 in such a manner that the phases or waveforms thereof are shifted by 180 degrees from each other. Therefore, the mixing diodes 302 and 303 have nonlinear conductances with the period of the local oscillator frequency. The incoming signal whose frequency is expressed in terms of f_R is applied via a band pass filter 312 to the mixing diodes 302 and 303 through a junction between the mixing diodes 302 and 303 as indicated by arrows in FIG. 6. Thus, the incoming signal frequency f_R is converted into an intermediate frequency f_{IF}, and the intermediate frequency developed at the junction is derived via a low pass filter 311.

Although the above-described conventional mixing circuits of FIG. 4 and FIG. 6 may be made small and compact by using a ferrite core for the transformer or transformers, the conversion loss assumes a large value, such as from 6 to 10 dB, since image frequency components and harmonic components of the local oscillator frequency are left as they are.

Generally speaking, when a local oscillator signal whose frequency is expressed in terms of f_L and an incoming or received signal whose frequency is expressed in terms of f_R are respectively applied to a mixing diode, signals corresponding to the sum and the difference (f_R-f_L and f_R+f_L) of and between the local oscillator frequency and the incoming signal frequency as well as the sum and the difference ($nf_L \pm f_R$) of and between harmonics of the local oscillator frequency and the incoming signal frequency are produced by the mixing diode (see FIG. 7). In the above, "n" is 2, 3, 4 . . . . Therefore, if an image frequency component $2f_L-f_R$ and a sum component f_L+f_R are effectively used by reconverting the same into the intermediate frequency f_R-f_L, the conversion loss of the mixing circuit would be considerably improved.

In the prior art, the image frequency component has been used in such a manner that the impedance viewed from the mixing diode equals zero or infinite by the use of an image frequency band pass filter and a band rejection filter for instance, in a waveguide type mixer and the balanced mixer of FIG. 5, which is made of an MIC, with a directional coupler. As a result, the conversion loss is improved to an extent. Furthermore, in some waveguide type mixers, a resonating window is provided in the waveguide for blocking the second harmonic of the local oscillator frequency and its side band signals.

However, in the prior art, an adequate consideration as to the sum component (f_L+f_R) of the local oscillator frequency and the incoming signal frequency as well as the harmonics of the local oscillator frequency, especially the second harmonic of the same, are not given.

Hereafter, the embodiments of the present invention will be described in detail with reference to FIG. 8 to FIG. 19 of the accompanying drawings.

Reference is now made to FIG. 8 which shows a schematic block diagram of a VHF-UHF all-band tuner for receiving Japanese TV channels according to the present invention. An input signal applied to an input terminal 40 is delivered to an input filter circuit 41. The filter circuit 41 comprises a trap circuit and a high pass filter for eliminating interference signals, which interfere with the intermediate frequency of a TV set, FM signals, and the like. This filter circuit 41 attenuates signals of the intermediate frequency and its half frequency of the TV set by approximately 40 dB. The output signal of the input filter circuit 41 is applied to a variable attenuator 42 which comprises a pin diode. The loss in the transmission band of the variable attenuator 42 is approximately 1 dB in the VHF band, and is approximately 1.5 dB in the UHF band, while the maximum amount of attenuation is approximately 50 dB in the VHF band, and is approximately 40 dB in the UHF band. The output signal of the variable attenuator 42 is applied to a radio frequency (RF) amplifier 43 which comprises two transistor stages. The gain of the RF amplifier 43 is approximately 21 dB, and the noise figure of the same is equal to or less than 2.5 dB in the VHF band, and is equal to or less than 3.5 dB in the UHF band. The output signal of the RF amplifier 43 is applied to a first input terminal of a first mixing circuit 45 which comprises a diode single balanced mixer. The first mixing circuit 45 has a second input terminal for receiving a local oscillator signal. Namely, a voltage controlled oscillator 47, which comprises a single transistor stage, generates the above-mentioned local oscillator signal where the output signal of the voltage controlled oscillator 47 is amplified by a local oscillator signal amplifier 46 to a given level before applied to the second input terminal of the first mixing circuit 45. The voltage controlled oscillator 47 generates a necessary oscillating signal corresponding to a desired TV channel by voltage controlling. The amplifier 46 amplifies the output signal of the voltage controlled oscillator 47 to obtain an output of +15 dBm throughout 2.5 to 3.5 GHz. The first mixing circuit 45 converts the TV signal applied to its first input terminal by mixing the same with the local oscillator signal applied via the amplifier 46 from the voltage controlled oscillator 47, into a given first intermediate frequency of 2520 to 2700 MHz. If 2520 MHz is selected as the first intermediate frequency, which does not give interference, and also if the mixing method of detecting the difference between a desired TV channel signal frequency and the local oscillator frequency, i.e. the output signal frequency of the voltage controlled oscillator 47, applied to the first mixing circuit 45, the necessary frequency range of the voltage controlled oscillator 47 is from 2613 to 3287 MHz.

The first mixing circuit 45 is a novel one according to the present invention, and is of the diode single balanced type having a low-noise characteristic and a low-distortion characteristic. The noise figure of the first mixing circuit 45 is 7 dB in the VHF band, and is 8 dB in the UHF band, while the conversion loss on frequency conversion to 2520 MHz is 6 dB in the VHF band, and is 7 dB in the UHF band. The output signal of the first mixing circuit 45 is applied to a fixed frequency band pass filter 48 which selectively transmits the TV channel signal whose frequency has been converted into the first intermediate frequency of 2520 MHz. The fixed frequency band pass filter 48 is of coaxial type, and the band width thereof is 5 MHz so that interferences by adjacent channels in the following circuits are prevented to the utmost. The insertion loss of the fixed frequency band pass filter 48 is approximately 5 dB at the center frequency of 2520 MHz. The output signal of the fixed frequency band pass filter 48 is applied to a first input terminal of a second mixing circuit 49 having a second input terminal for receiving the output signal of a fixed oscillator 50. If the first intermediate frequency has been selected as 2520 MHz, the output frequency of the fixed oscillator 50 is 2463 MHz when intended to receive Japanese TV channel signals. The

second mixing circuit 49 is of diode single balanced type, and the conversion loss thereof is 5 dB, and the noise figure thereof is 5 dB. The second mixing circuit 49 converts the first intermediate frequency, which has been selectively transmitted via the frequency fixed band pass filter 48, into a second intermediate frequency. In case of receiving Japanese TV channels the first intermediate frequency is converted into a band of 57 MHz.

The output signal of the second mixing circuit 49 is applied to an intermediate frequency (IF) amplifier 51 which amplifies the second intermediate frequency signal. The gain and the noise figure of the second intermediate frequency amplifier 51 are respectively 15 dB and 2.5 dB, while the level of the immediately adjacent channel signals, which give cross modulation distortion of 1 percent to the desired channel, is greater than −20 dBm. A reference numeral 52 indicates the combination of a trap and a fixed frequency band pass filter for attenuating the video carrier signal component and the sound carrier signal component of the upper and lower immediately adjacent channels so that interference by an adjacent channel signal in a following intermediate frequency amplifier is suppressed. The insertion loss of the above-mentioned combination of the trap and the frequency fixed band pass filter at the second intermediate frequency of 57 MHz is approximately 3 dB, while the amount of attenuation at the video carrier frequency of 52.75 MHz of the upper immediately adjacent channel and at the sound carrier frequency of 60.25 MHz of the lower immediately adjacent channel is approximately 30 dB. The output signal of the band pass filter 52 is applied to an amplifier 53 of output tuning type to be amplified to a given level. The gain of the amplifier 53 is 20 dB. In FIG. 8, a terminal 55 for supplying an AGC voltage is connected to the variable attenuator 42, while terminals 56 and 57 respectively for supplying a tuning voltage and an AFC voltage are connected to the voltage controlled oscillator 47.

The automatic gain control (AGC) in the tuner of FIG. 8 is adjusted in such a manner that the input level of the desired channel signal of the RF amplifier 43 is equal to or below −55 dBm.

Generally, in a tuner of double superheterodyne type, it is necessary to perform automatic frequency controls (AFC) with respect to the local oscillator, i.e. the voltage controlled oscillator, of the first mixing circuit, and to the fixed oscillator of the second mixing circuit. However, in accordance with the preferred embodiment of the present invention of FIG. 8, only the voltage controlled oscillator 47 is equipped with an AFC terminal because an oscillator, whose frequency is stable against the voltage variation in power source and against the temperature variation, is used as the fixed oscillator 50.

In accordance with the present invention the image frequency in connection with the first mixing circuit 45 equals a frequency which is higher than the desired channel frequency by twice the first intermediate frequency, i.e. 5040 MHz. Thus, in case of receiving Japanese TV band signals, the image frequency lies in a range from 5133 to 5807 MHz. Since the input terminal of the first mixing circuit 45 is connected to the output terminal of the RF amplifier 43, problem with respect to image interference in the first mixing circuit 45 does not arise at all.

The image frequency in connection with the second mixing circuit 49 is lower than the first intermediate frequency by twice the second intermediate frequency. Although only the fixed frequency band pass filter 48 is used to reject the image frequency, attenuation of 75 to 80 dB is obtained by means of the fixed frequency band pass filter 48 in a frequency band, which is lower than the center frequency by 114 MHz, namely in the vicinity of 2406 MHz so that the suppressing characteristic of the image signal which occurs in the second mixing circuit 49 is 75 to 80 dB.

In this embodiment, the total noise figure of the tuner is approximately 5 dB in the VHF band, and is approximately 7 dB in the UHF band. When the level of a desired channel signal is below −55 dBm, which is considered as the severest condition with respect to the interference characteristic of a tuner, and when the variable attenuator does not work, then the input signal level of the immediately adjacent or second adjacent channel signals, which give cross modulation distortion of 1 percent to the second intermediate frequency signal of the desired channel signal, is from −20 to −25 dBm.

Hereinbelow, characteristic individual circuits in the embodiment will be described in detail. FIG. 9 shows a schematic circuit arrangement of the first mixing circuit 45 of FIG. 8. This mixing circuit 45 has an original structure using microstriplines, and is of diode single balanced type. The mixing circuit 45 comprises microstriplines formed on a dielectric substrate (not shown), such as a plate made of Teflon (trademark) fiber glass, diodes, capacitors and resistors. The first mixing circuit 45 is an up-converter which converts the frequency of a TV channel signal of the VHF or UHF bands into an appropriate first intermediate frequency between 2520 and 2700 MHz, where the mixing circuit features that it is compact in size and it is of low-distortion characteristic.

The circuit arrangement of FIG. 9 is formed on a substrate made of a Teflon (trademark) fiber glass, and the size thereof is approximately 28 millimeters by 25 millimeters. In FIG. 9, reference numerals 61, 62 and 63 respectively designate terminals made of striplines of 50 ohms for respectively receiving an incoming signal, for deriving the first intermediate frequency signal, and for receiving the local oscillator signal. The incoming signal frequency is from 50 to 1000 MHz, while the local oscillator frequency is from 2570 to 3520 MHz.

The combination of a stripline 65 and striplines 66 and 67 constitute parallel coupled lines where the space between the parallel striplines 65, 66 and 67 is approximately 0.1 millimeter. Assuming that the wavelength of approximately the center frequency, i.e. 3000 MHz, of the local oscillator signal on the substrate is expressed in terms of λ_{gLL} the electrical length of the parallel coupled portion between the stripline 65 and the striplines 66 and 67 is approximately $\lambda_{gLL}/8$, and in this length, the electrical length of the parallel coupled portion between the striplines 65 and 66 is approximately $\lambda_{gLL}/40$.

A junction portion between the striplines 66 and 67 is connected via a ground stripline 77 to ground. An open-ended stub 64 is connected to the stripline 65, and the stub 64 has a characteristic impedance of 50 ohms and an electrical length of approximately $\lambda_{gLL}/10$. The parallel coupled lines 65, 66 and 67, the stub 64, and the ground stripline 77 constitute a balance-to-unbalance transformer (balun) at the local oscillator frequency band.

Reference numerals 68 and 69 respectively designate output striplines of the balun, and a chip resistor 71 and a chip capacitor 72 are connected to the output stripline

68, while a chip resistor 73 and a chip capacitor 74 are connected to the other output stripline 69. The resistances and the capacitances of these resistors 71 and 73, and the capacitances of these capacitors 72 and 74 are selected so that a low-distortion characteristic can be actualized without increasing the conversion loss. For instance, the resistances are respectively 10 ohms and the capacitances are respectively 5 picofarads. The resistor 71 and the capacitor 72 constitute a parallel circuit between the output stripline 68 and a first terminal 78, while the resistor 73 and the capacitor 74 constitute a parallel circuit between the output stripline 69 and a second terminal 79. An anode of a silicon Shottky barrier diode 70 is connected to the first terminal, while a cathode of another silicon Shottky barrier diode 70' is connected to the second terminal 79. The cathode of the Shottky barrier diode 70 and the anode of the other Shottky barrier diode 70' are connected to each other at a stripline 61'. A reference numeral 76 designates a low pass filter for receiving the incoming signal, from the terminal 61 while a reference numeral 75 designates a high pass filter through which the output intermediate frequency is delivered to the terminal 62. The output terminal of the low pass filter 76 and the input terminal of the high pass filter 75 are respectively connected to the above-mentioned stripline 61'. The combination of the high pass filter 75 and the low pass filter 76 constitutes a diplexer for separating input and output signals.

The mixing circuit of FIG. 9 actualizes a low-noise characteristic and a low-distortion characteristic by the provision of the balun for the local oscillator frequency, having an original structure and a balanced characteristic throughout a wide range, and by inserting parallel circuits of suitable resistors and capacitors between one output terminal 68 of the balun and one terminal of the diode 70 and also between the other output terminal 69 of the balun and one terminal of the other diode 70'.

Prior to describing the circuit arrangement of the second mixing circuit 49 of FIG. 8, the fundamental circuit arrangement for the second mixing circuit 49 will be described with reference to FIG. 10 for a better understanding of the structure and operation of the second mixing circuit 49.

FIG. 10 shows the fundamental structure of the second mixing circuit 49. A reference numeral 313 designates an input terminal for receiving a local oscillator signal f_L, and 314 designates an open-ended stub. Numerals 315, 316, 317, 318, 319, 336 and 337 are all striplines. The stripline 315 is connected at its one end to the input terminal 313, and is connected at its other end to one end of the stripline 336. The other end of the stripline 336 is connected to one end of the above-mentioned stub 314. The stripline 337 is arranged in parallel with the stripline 336 having a given space therebetween. Both ends of the stripline 337 are respectively connected to one ends of the striplines 318 and 319. A series circuit of mixing diodes 320 is interposed between the other ends of the striplines 318 and 319. A junction (no numeral) connecting the two diodes 320 is connected to a terminal 321. This terminal 321 is used for receiving the incoming signal f_R for deriving the intermediate frequency signal F_{IF}. The stripline 317 whose one end is connected to ground is connected at the other end thereof to the stripline 337 at about a midway point of the stripline 337. The junction point between the striplines 317 and 337 is designated as 338. In the same manner, the junction points between the striplines 315 and 336, between the stripline 336 and the stub 314, between the striplines 318 and 337, and between the striplines 319 and 337 are respectively designated as 331, 333, 334 and 335.

The striplines 336 and 337, which are arranged in parallel, constitute parallel coupled lines, and a point on the stripline 336 corresponding to the junction 338 is designated as 332 for the sake of description. The combination of the coupling stripline defined between the point 332 and the junction 331 and the coupling stripline defined between the junctions 338 and 334 constitutes an in-phase transformer. On the contrary, the combination of the coupling stripline defined between the point 332 and the junction 333 and the coupling stripline defined between the junctions 338 and 335 constitutes an opposite-phase transformer.

The above-mentioned stripline 316 is connected between the coupling striplines which constitute the above-mentioned opposite phase transformer. With this arrangement, the whole coupling portion in the transmission lines constitute a balance-to-unbalance transformer for the local oscillator frequency band. Since the stripline 316 is provided to make a short circuit between the striplines 336 and 337, the impedance viewed from the transmission lines 318 and 319, to which the mixing diodes 20 are respectively connected, equals zero at the sum component frequency band of the local oscillator frequency f_L and the incoming signal frequency f_R as well as at the second harmonic frequency $2f_L$ band of the local oscillator frequency.

In general, when comparing an in-phase or positive-phase transformer with an opposite-phase or negative-phase transformer, the latter has superior high frequency characteristics. This is because the resonance condition as to the length of the coupled lines is established in the vicinity of $\beta l = \pi/2$ in case of the in-phase transformer, and in the vicinity of $\beta l = \pi$ in case of the opposite-phase transformer wherein β is the phase constant, and l is the length of the coupled lines.

Accordingly, the coupling coefficient in the in-phase transformer at the frequency bands of f_L+f_R and $2f_L$ is small, whereas the coupling coefficient in the opposite-phase transformer at the same frequency bands is great. As a result, balance between the in-phase and opposite-phase transformers is lost so that the opposite-phase transformer side of the entire transformer has a load Z_L at the above-mentioned frequency bands as shown in an equivalent circuit of FIG. 11A where the load Z_L varies in accordance with the coupling degree of the opposite-phase transformer. The in-phase transformer side has an impedance substantially equal to zero as the load thereof because the coupling degree is extremely small.

In accordance with the present invention the above-mentioned stripline 316 for making a short circuit between the striplines 336 and 337 is additionally employed to make the impedances viewed from the striplines, to which the mixing diodes 320 are respectively connected, substantially zero as illustrated in a second equivalent circuit of FIG. 11B. In order that the transformer has a characteristics represented by the equivalent circuit of FIG. 11B the position of the stripline 316 has to be selected in view of the characteristic impedance of the open-ended stub 314 and the electrical length of the same.

However, the position of the stripline 316 for making the short circuit may also be selected in accordance with the following experimental method. Namely, a suitable position may be selected by detecting the minimum conversion loss while making a connection be-

tween the striplines 336 and 337 in the opposite-phase transformer by means of a suitable tool, such as a pair of tweezers.

With the above-mentioned arrangement, the sum component (f_L+f_R) among various signals produced by the mixing diodes 320 is almost perfectly reflected because of the zero impedance of the striplines 318 and 319 to which the mixing diodes 320 are respectively connected. As a result of the reflection, the sum component (f_L+f_R) is fed to the mixing diodes 320 once more to effect a frequency conversion expressed by $f_L+f_R-2f_L=f_R-f_L$. Consequently, the conversion loss characteristic can remarkably improved.

Hence, reference is now made to FIG. 12 which shows a schematic circuit arrangement of the second mixing circuit 49 of FIG. 8. The mixing circuit 49 has an original structure of diode single palance type where the mixing circuit 49 is based on the fundamental structure of FIG. 10. The second mixing circuit 49 has features that the circuit is compact in size and has a low-noise characteristic. The circuit arrangement of FIG. 12 is formed on a substrate (not shown) made of Teflon (trademark) fiber glass, and has a size of approximately 25 by 25 millimeters. In FIG. 12, reference numerals 81, 82 and 83 respectively designate terminals made of microstriplines of 50 ohms for respectively receiving an input signal, for deriving the second intermediate frequency signal, and for receiving the local oscillator signal. The input signal corresponds to the first intermediate frequency signal which has been selectively transmitted through the fixed frequency band pass filter 48, and the frquency assumes a suitable value between 2520 and 2700 MHz. In the embodiment of FIG. 8, the first intermediate frequency is 2520 MHz. The local oscillator frequency is lower than the input signal frequency, i.e. the first intermediate frequency, by the second intermediate frequency so that the necessary second intermediate frequency will be obtained. In the embodiment of FIG. 8, the local oscillator frequency is set to 2463 MHz when it is intended to receive Japanese TV band signals.

The combination of a stripline 85 and striplines 86 and 87 constitutes parallel coupled lines where the space between the parallel striplines 85, 86 and 87 is approximately from 0.06 to 0.1 millimeter. Assuming that the wavelength of the local oscillator frequency of 2463 MHz on the substrate is expressed in terms of λ_{gL2}, the electrical length of the parallel coupled portion along the stripline 85 and the striplines 86 and 87 is approximately $\lambda_{gL2}/10$, and in this length, the electrical length of the parallel coupled portion along the striplines 85 and 86, which constitute the in-phase transformer, is approximately from $\lambda_{gL2}/30$ to $\lambda_{gL2}/35$. The width of the striplines 85, 86 and 87 may be determined by various factors, such as the material of the substrate, and in this embodiment, the width is 0.4 millimeters.

A junction portion between the striplines 86 and 87 is connected via a ground stripline 93 to ground. A stripline 94 is connected between the stripline 85 and the stripline 86 at a midway point of the stripline 86 for making a short circuit therebetween. The stripline 94 will be referred to as a short-circuit stripline hereinafter. With the provision of the short-circuit stripline 94, the sum component $(f_{L2}+F_{IF1})$ of the local oscillator frequency (f_{L2}) and the input signal, i.e. the first intermediate frequency (f_{IF1}) can be effectively used to obtain the necessary difference component $(f_{IF1}-f_{L2})$ so that the conversion loss characteristic and the noise figure characteristic with respect to the second intermediate frequency are respectively improved. An open-ended stub 84 which corresponds to the stub 314 of FIG. 10, is connected to the stripline 85, and the stub 85 has a characteristic impedance of 35 ohms and an electrical length of approximately $\lambda_{gL2}/12$. The parallel coupled transmission lines 85, 86 and 87, the stub 84, the ground stripline 93, and the short-circuit stripline 94 constitute the above-mentioned balance-to-unbalance transformer (balun) at the local oscillator frequency. The balun has two output terminals 88 and 89 which respectively comprise striplines respectively connected to the striplines 87 and 86. An anode of a silicon Shottky barrier diode 90 is connected to the output terminal 88 and the cathode of the diode 90 is connected to a stripline 82'. In the same manner another silicon Shottky barrier diode 90' is connected between the output terminal 89 and the stripline 82' in such a direction that the anode thereof is connected to the terminal 89.

A reference numeral 92 designates a band pass filter for receiving the input signal from the terminal 81, while a reference numeral 91 designates a low pass filter through which the output second intermediate frequency signal is delivered to the terminal 82. The output terminal of the band pass filter 92 and the input terminal of the low pass filter 91 are respectively connected to the above-mentioned stripline 82'. The combination of the band pass filter 92 and the low pass filter 91 constitutes a diplexer for separating the input and output signals.

FIG. 13 is a graphical representation of the conversion loss in the above-described second mixing circuit 49. The conversion loss at a frequency of 2.5 GHz is 3.4 dB which is considerably smaller than a value obtained in the prior art.

FIG. 14A shows circuit arrangements of the voltage controlled oscillator 47 and the amplifier 46 shown in FIG. 8. In the voltage controlled oscillator 47, an inter-digital capacitor is used for coupling the resonator and the oscillating circuit thereof so that the voltage controlled oscillator 47 is capable of oscillating throughout a wide range, while the coupling portion does not require any adjustment. The voltage controlled oscillator 47 has an oscillation frequency band width of 1000 MHz so that it is possible to receive TV signals of VHF and UHF bands by means of a single tuner without effecting switching.

The circuits of FIG. 14A are formed on a substrate made of Teflon (trademark) fiber glass where the resonator is constructed of a stripline 99 and a varactor diode 100. The stripline 99 is of substantially U-shape so that the resonator can be small in size without deteriorating its characteristics. The oscillating frequency will be controlled by a voltage applied to a terminal 101.

A varactor diode 100 is connected between one side of the U-shaped stripline 99 and ground, while a resistor R79 is connected between the one side of the stripline 99 and the terminal 101 which is grounded via a capacitor C77. The other side of the U-shaped stripline 99 is connected to one terminal of the inter-digital capacitor 98. The other terminal of the inter-digital capacitor 98 is connected to a stub 97 which is employed for effecting impedance matching between the oscillating circuit side and the resonator. The stub 97 is connected to a stripline 96A which is connected via a series circuit of a resistor R78 and a choke coil L74 to ground where a junction connecting the resistor R78 and the choke coil L74 is connected via a capacitor C76 to ground. The opposite

end of the stripline 96A is connected to an emitter of a transistor (no numeral) the base of which is connected to a stripline 96B. A parallel circuit of a resistor R77 and a capacitor C78 is connected between the stripline 96B and ground, while a resistor R76 is connected between the stripline 96B and a terminal to which a power supply Vcc is connected. The collector of the transistor is connected to a stripline 96C which is ground via a capacitor C74. The stripline 96C is further connected via a series circuit of a choke coil L73 and a resistor R75 to the power supply terminal. A junction between the choke coil L73 and the resistor R75 is connected via a capacitor C75 to ground. The stripline 96C is further connected via a capacitor C73 to a base of a first transistor included in the amplifier 46. Each of the above-mentioned choke coils L73 and L74 has an electrical length expressed in terms of $\lambda_{gL1}/4$ assuming that the wavelength of the center frequency, i.e. 3000 MHz of the oscillator frequency on the substrate is expressed in terms of λ_{gL1}. The capacitors C75, C76, and C77 are used for transmitting high frequency signals to ground. With the provision of the combination of the stripline 96B and the capacitors C74 and C78 the negative resistance of the oscillating circuit can be controlled and in the preferred embodiment, the capacitance of the capacitor C74 is approximately 1 picofarad, and the capacitance of the capacitor C78 is between 5 and 10 picofarads. The characteristic impedance of these striplines 96A and 96C is 50 ohms, while the electrical length of these striplines 96A to 96C is preferably less than $\lambda_{gL1}/8$.

The local oscillator signal amplifier 46 comprises the above-mentioned first transistor (no numeral) and a second transistor (no numeral). A series circuit of resistors R73 and R74 is interposed between a power supply terminal and the base of the transistor. A coil L72 is interposed between a junction J1 connecting the resistors R73 and R74 and the collector of the first transistor the emitter of which is connected to ground. The collector of the first transistor is connected via a capacitor C72 to a base of the second transistor the emitter of which is grounded. A series circuit of resistors R71 and R72 is interposed between the power supply terminal and the base of the second transistor. A coil L71 is interposed between the collector and a junction J2 connecting the resistors R71 and R72. The collector of the second transistor is connected via a capacitor C71 to an output terminal 95 which is connected, inturn, to the second input terminal of the first mixing circuit 45 of FIG. 8. Two capacitors C113 and C114 are respectively interposed between the junctions J1 and J2 and ground.

With the provision of the combination of the voltage controlled oscillator 47 and the amplifier 46 it is possible to supply the first mixing circuit 45 with a local oscillator signal having an output power of 15 dBm throughout 2500 to 3500 MHz.

The voltage controlled oscillator 47 of FIG. 14A is equipped with an AFC (automatic frequency control) circuit. The AFC circuit of FIG. 14A comprises two transistors which are coupled in parallel between a power supply Vcc and ground. A base of a first transistor (no numeral) is connected via a resistor R82 to the terminal 56, while a base of a second transistor (no numeral) is connected via a resistor R83 to the terminal 57. As described with reference to FIG. 8, a variable voltage for effecting tuning is applied to the terminal 56, while a variable AFC voltage, which varies in accordance with the deviation in the oscillating frequency, is applied to the terminal 57. These two voltages are superimposed on each other to make a resultant voltage which is derived from an output terminal 102 connected via resistors R84 and R85 to the emitters of the first and second transistors. The output terminal 102 is connected to the terminal 101 of the voltage controlled oscillator 47 to supply the voltage controlled oscillator 47 with the resultant voltage produced in the AFC circuit. Since the oscillator frequency deviation is compensated for by means of this AFC circuit, a high stability in frequency, which is required when receiving color picture signals, is readily obtained. Although an undesirable phenomenon such that a current flows into the power supply side when the tuning voltage has a specific relationship with respect to the AFC voltage, is inevitable in an AFC circuit utilizing a voltage dividing circuit having resistors, such an undesirable phenomenon does not occur in the above-described AFC circuit of FIG. 14A.

FIG. 14B shows a schematic circuit diagram of a modified AFC circuit which may be used in combination with the voltage controlled oscillator 47 in place of the abovedescribed AFC circuit of FIG. 14A. In this modified AFC circuit, the AFC voltage is applied via the terminal 57 to an additionally provided varactor 400 connected to the U-shaped resonator 99 of the voltage controlled oscillator 47, while the tuning voltage is applied to the tuning varactor 100 directly via the terminal 56 which corresponds to the terminal 101 of FIG. 14A.

In detail, the AFC circuit of FIG. 14B comprises a series circuit of a resistor R111 connected to the terminal 57 and a capacitor C112 connected to the U-shaped resonator 99 in the vicinity of the inter-digital capacitor 98. The above-mentioned additionally provided varactor 400 is connected between a junction connecting the resistor R111 and the capacitor C112, and ground. A capacitor C111 is connected between the terminal 57 and ground.

By the use of the simple AFC circuit of FIG. 14B there is no need to use the AFC circuit of FIG. 14A. The AFC circuit of FIG. 14B has an advantage that the AFC response is uniform throughout the entire variable frequency range. In other words, the degree of the frequency variation by a given AFC voltage is substantially constant throughout the wide local oscillator frequency band. Since the additionally provided varactor 400 is used only for AFC, the maximum variable frequency range is not required to be as large as that of the tuning varactor diode 100. Therefore, a varactor of low cost may be used as the varactor 400.

Reference is now made to FIG. 15 which shows a circuit diagram of the fixed frequency oscillator 50 of FIG. 8. The fixed oscillator comprises a substantially U-shaped half wave resonator equipped with a loading capacitance. Therefore, the oscillator can be compact in size without deteriorating the characteristics thereof. A magnetic coupling is adopted in order to couple the resonator with an oscillating circuit in the fixed oscillator, where the coupling coefficient varies in accordance with the distance from an apparent short-circuited point at the center of the half wave resonator. Furthermore, in the oscillator of FIG. 15, the coupling between the resonator and the oscillating circuit is made as loose as possible, while biasing conditions of the transistor and the emitter resistor are suitably selected so that a very stable oscillator is actualized with a small number

of parts. In detail, the frequency variation can be suppressed within 200 to 300 KHz.

The fixed frequency oscillator of FIG. 15 is formed on a substrate made of Teflon (trademark) fiber glass. The above-mentioned resonator comprises a U-shaped stripline 107 and a fixed capacitor C95 connected to one end of the U-shaped stripline 107. The U-shaped stripline 107 is connected via a capacitor C94 to a stripline 105 connected to the emitter of the transistor, while the stripline 105 is connected via a resistor R96 to ground. The base of the transistor is connected to a stripline 106 coupled via a parallel circuit of a capacitor C93 and a resistor R95 to ground. The stripline 106 is connected via a resistor R94 to a power supply Vcc terminal. The collector of the transistor is connected to a stripline 104 connected via a capacitor C92 to ground. A series circuit of a choke coil L90 and a resistor R93 is interposed between the stripline 104 and the terminal connected to the power supply Vcc. A junction connecting the choke coil L90 and the resistor R93 is connected via a capacitor C91 to ground. The stripline 104 is connected via a series circuit of a capacitor C90 and a resistor R91 to an output terminal 103 which is grounded via a resistor R90. A resistor R92 is interposed between a junction connecting the capacitor C90 and the resistor R91, and ground. The characteristic impedance of these striplines 104 and 105 is 50 ohms, while the electrical length of these striplines 104 and 105 is preferably less than $\lambda_{gL2}/8$.

With this arrangement, an output oscillating frequency is developed at the output terminal 103 to be applied to the second input terminal of the second mixing circuit 49 of FIG. 8. The choke coil L90 has an electrical length expressed in terms of $\lambda_{gL2}/4$ assuming that the wavelength of the oscillating frequency of 2463 MHz on the substrate is expressed in terms of λ_{gL2}. The capacitor C91 is used for transmitting high frequency signals to ground. The negative resistance of the oscillating circuit can be controlled by means of the combination of the capacitors C92 and C93 and the stripline 106. In the preferred embodiment, the capacitance of the capacitor C92 is approximately 1 picofarad, and the capacitance of the capacitor C93 is between 10 and 20 picofarads. Fine adjustment of the oscillating frequency may be achieved by changing the space between a tuning adjust screw, which is located close to the open end of the U-shaped microstripline 107, and the open end.

FIG. 16 illustrates a schematic top plan view of the fixed frequency filter 48 of FIG. 8, which filter 48 is used for selectively transmitting the first intermediate frequency signal. The filter 48 is of three-stage resonator type where coupling between adjacent stages is effected by means of coupling through-holes made in shielding plates.

The filter of FIG. 9 has a size expressed in terms of 26 (width)×40 (length)×14 (height) millimeters. The filter 48 comprises a casing 114, three resonators 109, 110 and 111, and adjust screws 108. The combination of the resonators 109 to 111 and the adjust screws 108, by which tuning is effected, constitutes a resonating circuit. Two strip-like dielectric substrates 116 are fixedly connected respectively by means of screws 117 to two resonators 109 and 111 which are located at both ends in the casing 114. Two input and output transmission lines 115 are respectively placed on the dielectric substrates 116 where one end of each of the transmission lines 115 projects outwardly beyond the casing 114. These input and output transmission lines 115 are used for directly coupling the filter between two microwave integrated circuits corresponding to the first and second mixing circuits 45 and 49 of FIG. 8. Since coaxial connectors are not used for coupling the filter, the filter does not suffer from impedance miss matching so that characteristic deterioration due to coupling is prevented. In addition, since inter-stage coupling is effected through the coupling through-holes 113 made in the shielding plates 112, characteristics of high harmonic spurious signals of the filter are improved so that spurious signal interference due to mixing of harmonics of the two local oscillator signals from both the voltage controlled oscillator 47 and the fixed frequency oscillator 50 is reduced. The band width of the filter 48 is set between 3 and 5 MHz in order to ensure the characteristic of the desired channel signal and to improve the interference-rejection characteristics in the following circuits. The characteristics of the filter 48 assuming that the center frequency is 2520 MHz and the band with is 5 MHz, are such that the insertion loss is less than 5 dB; input/output VSWR is below 1.5; and the attenuation is 75 to 80 dB at frequencies expressed in terms of the center frequency ±114 MHz.

FIG. 17 is a graphical representation showing the relationship between the frequency response of the frequency fixed band pass filter 48 and the frequency response of the second intermediate frequency amplifying stage 51A, i.e., the combination of the intermediate frequency amplifier 51, the fixed frequency band pass filter 52 and the amplifier 53. In FIG. 17 a reference numeral 190 indicates the former, and 191 indicates the latter. The frequency responses illustrated by curves respectively indicate the relative attenuation with respect to the attenuation at the center frequencies thereof. A frequency scale designated as f_1 corresponds to the frequencies for the response curve 190, while another frequency scale designated as f_2 corresponds to the frequencies for the response curve 191. References N, N+1, N−1, N+2 and N−2 respectively indicate the desired channel, an upper immediately adjacent channel, a lower immediately adjacent channel, an upper second adjacent channel and a lower second adjacent channel. Arrows respectively shown at the right side of the channels indicate the video carrier frequency side, and straight line at the left sides of channels indicate the sound carrier frequency sides. As illustrated in the graph of FIG. 17, the attenuation degree of the filter 48 at the second adjacent channels (N+2 and N−2) is over 20 dB (see curve 190). Although the attenuation degrees of the filter 48 at the video carrier frequency of the lower immediately adjacent channel N−1 and at the sound carrier frequency of the upper immediately adjacent channel N+1 are respectively over 15 1 dB (see points A and B), the attenuation degrees of the same at the sound carrier frequency of the lower immediately adjacent channel N−1 and at the video carrier frequency of the upper immediately adjacent channel N+1 are respectively several dB (see points C and D).

Since some signals in the immediately adjacent channels N+1 and N−1 readily transmit through the band pass filter 48, these transmitted signals are apt to cause interferences in the following stages if no countermeasure is taken. However, in accordance with the present invention the combination of the intermediate frequency amplifier 51 and the following circuits 52 and 53 has the above-mentioned frequency response which is

represented by the curve 191 so that interference-rejection characteristic is improved.

FIG. 18 shows a circuit diagram of the above-mentioned second intermediate frequency amplifying stage 51A including the intermediate frequency amplifier 51, the band pass filter 52 and the output intermediate frequency amplifier 53. Reference numerals 192 and 193 respectively designate the input and output terminals of the amplifying stage 51A. The intermediate frequency amplifier 51 comprises a transistor the emitter of which is connected via an input coupling capacitor C101 to the input terminal 192. The collector of the transistor is connected via an output coupling capacitor C104 to an input of the following band pass filter 52, and is further connected via a resistor to a terminal connected to a power supply Vcc. A series circuit of resistors R101 and R102 constitutes a voltage divider between the power supply Vcc and ground, and a junction between these resistors R101 and R102 is connected to the base of the transistor for serving as a bias circuit.

A parallel circuit of a resistor R104 and a capacitor C102 is connected at its one end to the emitter of the transistor, and is further connected at the other end thereof via another parallel circuit of a resistor R103 and a capacitor C103 to ground. Namely, a parallel-series circuit of resistors R104 and R103 and capacitors C102 and C103 is interposed in the grounded-emitter circuit. This parallel-series circuit is utilized for obtaining low-noise and low-distortion characteristics.

The capacitor C102 connected across the resistor R103 is used for transmitting the intermediate frequency signal to ground. The capacitance and resistance of the capacitor C102 and resistor R104 are respectively selected so as to actualize the low-distortion characteristic without causing the low-noise characteristic of the intermediate frequency amplifier 51 to be deteriorated. In the preferred embodiment, the resistance of R104 is from 10 to 30 ohms, while the capacitance of the C102 is from 100 to 200 picofarads.

With the provision of the resistor R104 and the capacitor C102 the cross modulation distortion characteristic at the second intermediate frequency band is considerably improved. If the resistor R104 and the capacitor C102 were not added, namely if only the parallel circuit of the resistor R103 and the capacitor of C103 were interposed between the emitter and ground, the noise figure would be 2.0 dB, while the input level of adjacent channel signals, which give cross modulation distortion of 1 percent to the desired channel signal, would be −30 dBm.

In the embodiment of FIG. 18, however, the noise figure of the intermediate frequency amplifier 51 is below 2.5 dB, and the input level of adjacent channel signals, which give cross modulation distortion of 1 percent to the desired channel signal, is over −20 dBm owing to the provision of the resistor R104 and the capacitor C102.

The band pass filter 52 is fundamentally constructed of a ceramic filter P_2 and ceramic resonators P_1 and P_3. The ceramic filter P_2 is of one-resonator type, and a series circuit of a capacitor C105 and a coil L102 is connected to its ground terminal. The ceramic resonator P_1 functions as a trap filter at the sound carrier frequency, i.e. 60.25 MHz, of the lower immediately adjacent channel N−1. The ceramic resonator P_3 functions as a trap filter at the video carrier frequency, 52.75 Mhz, of the upper immediately adjacent channel. The attenuation degree of each trap is over 30 dB. Two coils L101 and L103 are used for effecting impedance matching at the input and output of the band pass filter 52.

The output intermediate frequency amplifier 53 comprises a tuning amplifier having a resonating circuit constructed of a capacitor C109 and a coil L104. The coil L104 is inductively coupling with a coil L105 in such a manner that the output impedance at 57 MHz band is 75 ohms.

With the provision of the combination of the second intermediate frequency amplifier 51, the band pass filter 52, and the output intermediate frequency amplifier 53 a noise figure of below 3 dB is obtained, while the input level of immediate adjacent channel signals, which give cross modulation distortion of 1 percent to the desired channel signal, is over −20 dBm.

Reference is now made to FIG. 19 which shows a schematic block diagram of the second embodiment of the VHF-UHF all band tuner according to the present invention. The tuner of FIG. 19 is for receiving TV channel in the United States of America. The second embodiment has better characteristics with respect to two-channel beat interference, second harmonic interference and cross modulation interference than the first embodiment of FIG. 8. Elements and circuits which are the same as those in FIG. 8 are designated at like numerals, whereas elements and circuits whose set frequency differs from those in FIG. 8 are designated at like numerals which are apostrophized.

Reference numerals 194 and 195 respectively designate a VHF input terminal and a UHF input terminal. A band pass filter 196 for VHF lower channels having a band from 50 to 90 MHz and a band pass filter 197 for VHF higher channels having a band from 170 to 220 MHz are respectively connected to the VHF input terminal 194. The output terminals of the band pass filters 196 and 197 are connected to input terminals of a switching circuit 199. The switching circuit 199 selectively transmits one of the output signals of the band pass filters 196 and 197 in accordance with a control voltage applied to a terminal 205 where the control voltage is variable to a desired channel. The output of the switching circuit 199 is connected to an input terminal of a variable attenuator 421 which is constructed of a pin diode. The insertion loss of each of the band pass filters 196 and 197 and the switching circuit 199 is below 1 dB. The variable attenuator is the same in construction as that of the first embodiment, and the loss at the transmission band is below 1 dB while the maximum attenuation degree is 50 dB. The output signal of the variable attenuator 421 is applied to a radio frequency (RF) amplifier 200 of the VHF band. The RF amplifier 200 is a single-transistor stage amplifier having a gain of 15 dB and a noise figure below 2.5 dB at a band from 50 to 250 MHz. The input level of the immediately adjacent channels and second adjacent channels, which give cross modulation distortion of 1 percent to the desired channel signal, is below approximately −15 dBm.

A band pass filter 198 for the UHF band is connected to the UHF input terminal 195 where the band pass filter 198 has a band from 450 to 900 MHz. The output terminal of the band pass filter 198 is connected to a variable attenuator 422 which is constructed of a pin diode. The loss at the UHF transmission band of the variable attenuator 422 is below 1.5 dB while the maximum attenuation degree is 45 dB. The output signal of the variable attenuator 422 is applied to an input terminal of an RF amplifier 201 of UHF band having two transistor amplifying stages. The RF amplifier 201 has a

gain of 15 dB and a noise figure below 3.5 dB at a band from 400 to 1000 MHz, and the input level of immediately adjacent channels and second adjacent channels, which give cross modulation distortion of 1 percent to the desired channel signal, is below approximately −15 dBm. Both the output terminals of the VHF RF amplifier 200 and the UHF RF amplifier 201 are respectively connected to input terminals of a VHF/UHF switching circuit 202 which is responsive to a control voltage applied to a terminal 206. Namely, the VHF/UHF switching circuit 202 selectively transmits one of the output signals of the RF amplifiers 200 and 201 to a following first mixing circuit 45' which is constructed of a diode single balanced mixer. The first mixing circuit 45' has a second input terminal for receiving a local oscillator signal which is generated by a voltage controlled oscillator 47' and is amplified by an amplifier 46' in the same manner as in the first embodiment.

Assuming that a frequency of 2700 MHz is selected as the first intermediate frequency, which does not give interference to the desired channel signals in the United States of America, and the first intermediate frequency equals the difference between the oscillating frequency of the voltage controlled oscillator 47' and the desired channel signal frequency, the necessary oscillating frequency band of the voltage controlled oscillator 47' is from 2757 to 3587 MHz.

The first mixing circuit 45', the voltage controlled oscillator 47' and the amplifier 46' are substantially the same in construction as those in the first embodiment where only the set frequency are slightly different from those in the first embodiment. The output signal of the first mixing circuit 45' is applied to a first amplifier 203 for the first intermediate frequency, which amplifier 203 is of single-transistor stage type. The first amplifier 203 has a gain of 6 dB and a noise figure below 3.5 dB at 2700 MHz, and the input level of immediately adjacent channel signals and second adjacent channel signals, which give cross modulation distortion of 1 percent to the desired channel signal, is approximately −8 dBm.

The output signal of the first amplifier 203 for the first intermediate frequency is applied to a fixed frequency band pass filter 48' of coaxial type having a band width of 5 MHz and a center frequency of 2700 MHz. The insertion loss of the band pass filter 48' at the center frequency is approximately 5 dB. The output signal of the frequency fixed band pass filter 48' is applied to a second amplifier 204 for the first intermediate frequency. The second amplifier 204 is of single-transistor stage, and has a gain of 6 dB and a noise figure over 3.0 dB at 2700 MHz. The output signal of the second amplifier 204 is applied to a first input terminal of a second mixing circuit 49' having second input terminal for receiving a local oscillator frequency signal of 2655 MHz generated by a fixed frequency oscillator 50'. The output signal of the second mixing circuit 49' is applied to a combination 52' of a frequency fixed band pass filter and a trap circuit for selectively transmitting the second intermediate frequency signal of 45 MHz. The circuit 52' is the same in construction as the circuit 52 of the first embodiment whereas the set frequency differs. The output signal of the filter circuit 52' is applied to an amplifier 53' of output tuning type to be amplified to a given level. The gain of the amplifier 53' is 25 dB.

Terminals 55, 56 and 57 for respectively receiving an AGC voltage, a tuning voltage, and an AFC voltage are provided in the same manner as in the first embodiment. Since the operations of AGC, tuning and AFC are the same as in the first embodiment, description thereof is omitted.

In the above-described second embodiment, since the first and second amplifiers 203 and 204 of low gain, low distortion and low noise type for the first intermediate frequency are respectively coupled to the input and output sides of the fixed frequency band pass filter 48', it is possible to lower the gain of the RF amplifiers 200 and 201 without deteriorating the total noise characteristics of the whole tuner. Accordingly, the amplifying gain between the input terminal 194 or 195 and the first input terminal of the first mixing circuit 45' can be reduced compared to the first embodiment so that the characteristics against cross modulation interference and inter-third harmonic interference in the whole tuner is improved when compared to the first embodiment. In accordance with the second embodiment the total noise figure of the tuner is approximately 5.5 dB at the VHF band, and is approximately 7 dB at the UHF band, while the input level of the immediately adjacent channel signals and the second adjacent channel signals, which give cross modulation distortion of 1 percent to the desired channel signal, is from −15 to 20 dBm.

Furthermore, with the provision of the band pass filters 196, 197 and 198 for VHF lower channels, VHF higher channels and UHF channels at the input side of each of the RF amplifiers 200 and 201, signals other than the desired channel singnal are attenuated before applied to the RF amplifier 200 or 201 so that the second harmonic interference rejection characteristic is improved. In addition, two-channel beat interference, which occurs when signals of two channels other than the desired channel are used to make the sum thereof and/or the difference therebetween to fall in the desired channel, is reduced.

As described hereinbefore, assuming that 2700 MHz is selected as the first intermediate frequency for receiving U.S. TV channels, and when the first intermediate frequency is obtained from the difference between the local oscillator frequency applied to the first mixing circuit 45' and the desired channel signal frequency, the spurious interferences which fall in the first intermediate frequency band are the second and third harmonic interferences which occur in the RF amplifier 200 or 201 and in the first mixing circuit 45', and beat interferences between channels. The rejection characteristic against these interferences has been remarkably improved in the second embodiment to the utmost.

Furthermore, in the second embodiment, the image frequency suppression characteristic in the VHF band, which image frequency occurs in the second mixing circuit 45', is over 85 dB because the image frequency is out of the bands of the band pass filters 196 and 197 for VHF lower channels and for VHF upper channels, also because of the provision of the frequency fixed filter 48'.

From the foregoing, it will be understood that the present invention provides a useful mixing circuit and VHF-UHF all band tuner having the mixing circuit. The above-described embodiment are just examples and therefore, many variations and modifications may be made without departing from the spirit of the present invention.

What is claimed is:

1. A VHF-UHF all-band tuner of the double superheterodyne type including a first mixing circuit for converting an incoming signal frequency into a first intermediate frequency by mixing said incoming signal with a first local oscillator signal whose frequency is

variable, a filter circuit for selectively receiving said first intermediate frequency signal, and a second mixing circuit responsive to the output signal of said filter circuit and to a second local oscillator signal whose frequency is fixed, for converting said first intermediate frequency into a second intermediate frequency; characterized in that said first local oscillator frequency is higher than said first intermediate frequency, and in that said first intermediate frequency lies in a band of 2520 to 2700 MHz.

2. A VHF-UHF all-band tuner as claimed in claim 1, wherein said first mixing circuit comprises a diode balanced mixer made in the form of a microwave integrated circuit.

3. A VHF-UHF all-band tuner as claimed in claim 2, wherein said diode balanced mixer comprises a balance-to unbalance transformer having first and second coupling lines which constitute parallel coupled lines, an open-ended stub connected to one end of said first coupling line, a line connected to the other end of said first coupling line for receiving said first local oscillator signal; and a ground line connected between said second coupling line at a substantially midway point of said second coupling line, and ground.

4. A VHF-UHF all-band tuner as claimed in claim 3, wherein said diode balanced mixer comprises first and second open-ended output lines respectively connected to both ends of said second coupling line; a terminal for receiving said incoming signal and for deriving said first intermediate frequency signal, said terminal being placed separately from said first and second open-ended output lines at a substantially midway point between said first and second open-ended output lines; a first series circuit, interposed between said first output line and said terminal, of a first resistor and a first mixing diode; a second series circuit, interposed between said second output line and said terminal, of a second resistor and second mixing diode; and first and second capacitors respectively connected across said first and second resistors.

5. A VHF-UHF all-band tuner as claimed in claim 4, further comprising a diplexer having a high pass filter connected to said terminal for receiving said incoming signal, and a low pass filter connected to said terminal for deriving said first intermediate frequency signal.

6. A VHF-UHF all-band tuber as claimed in claim 1, wherein said filter circuit has a band width of 3 to 5 MHz.

7. A VHF-UHF all-band tuner as claimed in claim 1, wherein said filter circuit comprises shielding plates respectively placed between resonators, each of said shielding plates having a coupling through-hole.

8. a VHF-UHF all-band tuner as claimed in claim 1, wherein said first local oscillator comprises a single tuning varactor by which the oscillating frequency is variable throughout an entire band corresponding to all channels of VHF to UHF bands.

9. A VHF-UHF all-band tuner as claimed in claim 8, wherein said first local oscillator comprises a voltage controlled oscillator having a resonator portion, oscillating circuit portion and an inter-digital capacitor made in the form of a microwave integrated circuit, said capacitor being placed between said resonator portion and said oscillating circuit portion for effecting coupling.

10. A VHF-UHF all-band tuner as claimed in claim 9, wherein said resonator comprises a substantially U-shaped stripline.

11. A VHF-UHF all-band tuner as claimed in claim 1, further comprising an AFC circuit in which an AFC control voltage is superimposed on a tuning voltage to produce a resultant voltage which is applied to said first local oscillator to change the oscillating frequency thereof.

12. A VHF-UHF all-band tuner as claimed in claim 9, further comprising an AFC circuit connected to said resonator portion, said AFC circuit having a varactor, the capacitance of which is variable in accordance with an AFC voltage.

13. A VHF-UHF all band tuner as claimed in claim 12, wherein said resonator portion comprises a stripline connected as its one end to said inter-digital capacitor, said AFC circuit being connected to said stripline close to said inter-digital capacitor.

14. A VHF-UHF all-band tuner as claimed in claim 13, wherein said AFC circuit comprises a series circuit of a resistor and a first capacitor interposed between a terminal for receiving an AFC voltage and said stripline; said varactor connected between a junction connecting said resistor and said first capacitor, and ground; and a second capacitor connected between said terminal and ground.

15. A VHF-UHF all-band tuner as claimed in claim 1, wherein said second local oscillator comprises a half wave resonator equipped with a loading capacitor, said half resonator being constructed to a microwave integrated circuit.

16. A VHF-UHF all-band tuner as claimed in claim 15, wherein said half wave resonator is connected to an oscillating circuit of second local oscillator at a substantially midway point of said stripline constituting said resonator.

17. A VHF-UHF all-band tuner as claimed in claim 1, wherein said second mixing circuit comprises a diode balanced mixer made in the form of a microwave integrated circuit.

18. A VHF-UHF all-band tuner as claimed in claim 17, wherein said diode balanced mixer comprises a balance-to-unbalance transformer having first and second coupling lines which constitute parallel coupled lines; an open-ended stub connected to one end of said first coupling line; a line connected to the other end of said first coupling line for receiving said second local oscillator signal; a ground line connected between said second coupling line at a substantially midway point of said second coupling line, and ground; and a short-circuit line connected between said first and second coupling lines.

19. A VHF-UHF all-band tuner as claimed in claim 18, wherein said diode balanced mixer comprises first and second open-ended output lines respectively connected to both ends of said second coupling line; a terminal for receiving said first intermediate frequency signal and for deriving said second intermediate frequency signal, said terminal being placed separately from said first and second open-ended output lines at a substantially midway point between said first and second open-ended output lines; and first and second mixing diodes respectively interposed between said first open-ended output line and said terminal, and between said second open-ended output line and said terminal.

20. A VHF-UHF all-band tuner as claimed in claim 19, further comprising a diplexer having a band pass filter connected to said terminal for receiving said first intermediate frequency signal; and a low pass filter

connected to said terminal for deriving said second intermediate frequency signal.

21. A VHF-UHF all-band tuner as claimed in claim 1, further comprising a second intermediate frequency amplifying stage having a second intermediate frequency amplifier and a band pass filter for selectively transmitting the second intermediate frequency from said second intermediate frequency amplifier.

22. A VHF-UHF all-band tuner as claimed in claim 21, further comprising a trap filter connected to said second intermediate frequency amplifier for attenuating a given frequency signal.

23. A VHF-UHF all-band tuner as claimed in claim 1, further comprising a low gain amplifier connected to the input side of said filter circuit.

24. A VHF-UHF all-band tuner as claimed in claim 1, further comprising an amplifier connected to the output side of said filter circuit.

25. A VHF-UHF all-band tuner as claimed in claim 1, further comprising two amplifiers respectively connected to the input and output sides of said filter circuit.

26. A VHF-UHF all-band tuner comprising:

(a) an input filter circuit responsive to an incoming or received signal, said input filter circuit having a trap circuit and a high pass filter for attenuating signals of the intermediate frequency and its half frequency of a TV set connected to said tuner;

(b) a variable attenuator responsive to the output signal of said filter circuit for attenuating the same in accordance with an AGC voltage;

(c) a radio frequency amplifier responsive to the output signal of said variable attenuator for amplifying the same to a given level;

(d) a voltage controlled oscillator for producing a first local oscillator signal whose frequency is variable in accordance with a tuning voltage and with an AGC voltage respectively applied thereto;

(e) a local oscillator signal amplifier responsive to the output signal of said voltage controlled oscillator for amplifying the same to a given level;

(f) a first mixing circuit responsive to the output signal of said radio frequency amplifier and to said local oscillator signal amplifier for converting the frequency of said incoming signal into a first intermediate frequency residing in a band ranging from 2520 to 2700 MHz by subtracting said incoming signal frequency from said first local oscillator frequency;

(g) a fixed frequency band pass filter for transmitting said first intermediate frequency signal from said first mixing circuit;

(h) a fixed frequency local oscillator for producing a second local oscillator signal whose frequency is constant;

(i) a second mixing circuit responsive to the output signal of said fixed frequency band pass filter and to said second local oscillator signal for converting said first intermediate frequency into a second intermediate frequency; and

(j) a second intermediate frequency amplifying stage responsive to said second intermediate frequency signal from said second mixing circuit, said second intermediate frequency amplifying stage having an intermediate frequency amplifier, a filter circuit constructed of a band pass filter and a trap circuit for attenuating the video carrier signal component of the upper immediately adjacent channel and the sound carrier signal component of the lower immediately adjacent channel both included in the output signal of said intermediate frequency amplifier, and an output intermediate frequency amplifier responsive to the output signal of said filter circuit.

27. A VHF-UHF all-band tuner comprising:

(a) a first input band pass filter responsive to an incoming or received signal of VHF lower channels;

(b) a second input band pass filter responsive to an incoming or received signal of VHF higher channels;

(c) a switching circuit for transmitting one of the output signals of said first and second input filters in accordance with a control voltage determined by a desired channel;

(d) a first variable attenuator responsive to the output signal of said switching circuit for attenuating the same in accordance with an AGC voltage;

(e) a first radio frequency amplifier responsive to the output signal of said variable attenuator for amplifying the same to a given level;

(f) a third input band pass filter responsive to an incoming signal of UHF channels;

(g) a second variable attenuator responsive to the output signal of said third input band pass filter for attenuating the same in accordance with said AGC voltage;

(h) a second radio frequency amplifier responsive to the output signal of said second variable attenuator for amplifying the same to a given level;

(i) a VHF/UHF switching circuit for transmitting one of the output signals of said first and second radio frequency amplifiers in accordance with a control voltage determined by a desired channel;

(j) a voltage controlled oscillator for producing a first local oscillator signal whose frequency is variable in accordance with a tuning voltage and with an AFC voltage respectively applied thereto;

(k) a local oscillator signal amplifier responsive to the output signal of said voltage controlled oscillator for amplifying the same to a given level;

(l) a first mixing circuit responsive to the output signal of said VHF/UHF switching circuit and to said local oscillator signal amplifier for converting the frequency of said incoming signal into a first intermediate frequency residing in a band from 2520 to 2700 MHz by subtracting said incoming signal frequency from said first local oscillator frequency;

(m) a first intermediate frequency amplifying stage for amplifying said first intermediate frequency signal from said first mixing circuit, said first intermediate frequency amplifying stage having a first amplifier of low gain; a band pass filter responsive to the output signal of said first amplifier; and a second amplifier of low gain responsive to the output signal of said band pass filter;

(n) a fixed frequently local oscillator for producing a second local oscillator signal whose frequency is constant;

(o) a second mixing circuit responsive to the output signal of said fixed frequency band pass filter and to said second local oscillator signal for converting said first intermediate frequency into a second intermediate frequency; and

(p) a second intermediate frequency amplifying stage responsive to said second intermediate frequency signal from said second mixing circuit, said second intermediate frequency amplifying stage having a

filter circuit constructed of a fixed frequency band pass filter and a trap circuit, and an amplifier.

28. A microwave mixing circuit comprising:

(a) a parallel coupled lines having first and second lines;

(b) an open-ended stub connected to one end of said first line;

(c) a first terminal connected to the other end of said first line, for receiving a local oscillator signal;

(d) a pair of mixing diodes connected in series between both ends of said second line;

(e) a ground line connected between said second line at its substantially midway point and ground;

(f) a second terminal connected between said pair of diodes for receiving an input signal whose frequency is to be converted, and for deriving an intermediate frequency signal; and

(g) a short-circuit line connected between said first and second lines.

29. A microwave mixing circuit as claimed in claim 28, further comprising a diplexer having a band pass filter through which said input signal is applied to said second terminal, and a low pass filter through which said intermediate frequency signal is derived from said second terminal.

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TAB 13

BERNARD J. LECHNER

Bernard J. Lechner is one of the world's leading experts on television and display systems. He is extensively involved in technical research on advanced television and display systems and in the development of standards for High Definition Television (HDTV). Currently consulting to government and industry on all aspects of television and display systems, Mr. Lechner's clients include major Fortune "500" corporations, as well as start-up companies in the television and display industries.

Formerly Staff Vice President, Advanced Video Systems, RCA Laboratories, Mr. Lechner's 30-year career at RCA covered all aspects of television and display research, from early work on home Video Tape Recorders in the late 1950's, extensive development of flat-panel matrix displays in the 1960's including pioneering efforts on active-matrix liquid crystal displays, advanced two-way Cable TV systems and pay-TV systems in the early 1970's, electronic tuning systems and CCD comb-filters for TV receivers in the mid-1970's, automated broadcast cameras and CCD broadcast cameras in the late 1970's and early 1980's, to HDTV in the mid-1980's. The National Academy of Television Arts and Sciences awarded Emmy's to two of the broadcast camera projects for which Mr. Lechner led the research team. Mr. Lechner received two RCA Laboratories Outstanding Achievement Awards and a David Sarnoff Team Award in Science.

He has been an active participant in the various groups working on standards for Advanced Television Systems in the U.S. including the FCC Advisory Committee on Advanced Television Service (ACATS), the SMPTE and the Advanced Television Systems Committee (ATSC). He is chairman of the ATSC Specialists Group on Digital TV Transport Standards, and he also chairs the CEA/NCTA Digital Standards Subcommittee relating to standards for "Cable-Ready" Digital Television receivers. During 1989 and 1990, Mr. Lechner served as a member of the U.S. delegation to the extraordinary and final meetings of the Comite Consultatif International des Radiocommunications (CCIR) in Geneva concerning international HDTV standards.

He was chairman of the Teletext Committee of the Electronic Industries Association from 1980 to 1986 and has been a member of the National Cable Television Association Engineering Committee since 1977. He served for 15 years as Chairman of the Advisory Commission for Electrical Engineering at Mercer County Community College and has been a member of the Board of Directors of Palisades Institute for Research Services since 1981.

Mr. Lechner is a Fellow of the IEEE, the Society for Information Display (SID) and the Society of Motion Picture and Television Engineers (SMPTE). He is a member of Tau Beta Pi, Eta Kappa Nu and Sigma Xi. In 1971 SID named him the first recipient of the Frances Rice Darne Award for his outstanding contributions to matrix displays. In 1972 he was elected to the SID Board of Directors and has since served as Treasurer, Secretary,

Vice President and President (1978-1980) of SID. In 1983 he was named the first recipient of the Beatrice Winner Award for his contributions to SID.

In 1996 Mr. Lechner was awarded the David Sarnoff Gold Medal by SMPTE for his many contributions to the technologies essential to today's television systems. In May 2000, Mr. Lechner was honored by the ATSC as the first recipient of the ATSC Outstanding Contributor Award. The award has become an ATSC annual award named in his honor. In November 2001 Mr. Lechner received the SMPTE Progress Medal Award. The Progress Medal is SMPTE'S highest honor and was presented to Mr. Lechner in recognition of his outstanding technical contributions to the progress of the engineering phases of the television industry.

Mr. Lechner holds a BSEE degree from Columbia University and has done graduate work at Princeton University and the Harvard School of Business.

Mr. Lechner holds ten U.S. patents and is widely published in the areas of displays and television systems.

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